IF FILTER OPTIMIZATION
FOR CW-FM RADAR

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THESIS

IF FILTER OPTIMIZATION

FOR CW-FM RADAR

by

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IF filter responses are investigated from the output of the narrow bandpass filter for various target ranges. The effects of the IF signal pulse width and the duty cycle on the IF filter output including the change of the sidelobe levels out of the filter are examined.

Optimum parameters are obtained from the experimental study and from an analytical solution.

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ACKNOWLEDGEMENT

To my wife Engin.

I. CW-FM RADAR AND IF FILTER

A. CW-FM RADAR

The continuous-wave frequency-modulated radar transmitter produces a CW signal of constant amplitude whose frequency varies in a sawtooth fashion as shown in Figure 1. The receiver picks up a part of the echo signal after a delay time of $\tau = 2R/c$. The dashed lines in Figure 1(a) represent the echo signal. If the echo signal is heterodyned with a portion of the transmitted signal, a beat note f_b is obtained as shown in Figure 1(b).

Note that for a given target two different beat frequencies are obtained. The beat frequency obtained during the later part of the transmitted frequency ramp is directly proportional to the target range and is given by

$$f_b = \frac{2R\Delta f}{cT}$$

The block diagram of a CW-FM radar with a sideband superheterodyne receiver is shown in Figure 2. It should be noted that separate antennas have been used for transmission and reception. The IF (intermediate frequency) signal produced by the local oscillator is mixed with a portion of the transmitted signal. The output of the mixer consists of the carrier, sidebands on either side of the carrier, and higher harmonics. A filter selects one of the sidebands as a reference signal and rejects the

carrier and the other sideband. The sideband filter must have a large enough bandwidth to pass the modulation, but not the carrier or the other sideband. When an echo signal is present and mixed with the output of the above filter in the receiver mixer, one obtains a signal the frequency of which is $f_{IF} + f_b$. This signal is amplified and fed into the balanced detector together with the signal f_{IF} coming from the local oscillator. The output of the detector contains only the beat frequencies.

$$f_b = \frac{2R}{c} \cdot \frac{\Delta f}{T} \qquad f_b' = \Delta f - f_b$$

The beat frequencies are amplified by a low-frequency amplifier and then applied to the frequency-measuring circuits.

The radar is normally designed so that $T > 2R_{max}/c$ where R_{max} is the maximum target range of interest. Therefore the duration of f_b is greater than the duration of f_b'. The later tone is therefore dropped by turning the amplifier off during the initial portion of the sweep. Therefore, f_b' will be ignored in the following discussion.

B. DOPPLER EFFECT

When an electromagnetic wave of frequency f_o is transmitted to a moving target, the received echo is shifted in frequency by an amount f_d according to the Doppler equation

$$f_d = -\frac{2f_o V_r}{c} = -\frac{2f_o}{c} \cdot \frac{dR}{dt} = -\frac{2V_r}{\lambda}$$

where V_r represents the relative radial component of velocity of the target, C the velocity of propagation and $\lambda = C/f_o$ the wavelength of the transmission. Thus the doppler frequency shift yields information about the velocity of the target. For a moving target we have, then,

$$f_b = \frac{2R\Delta f}{cT} - \frac{2f_o V_r}{c}$$

Solving this equation for target range we have

$$R = f_b \frac{cT}{2\Delta f} + V_r \frac{f_o T}{\Delta f}$$

If V_r is unknown and the range is calculated from the first term in the above expression, there is a range error, ΔR, given by

$$\Delta R = V_r \frac{f_o T}{\Delta f}$$

For surface targets this range error can generally be made small. For example [5], if f_o = 9 GHz, T = 380 microseconds and Δf = 12 MHz, the range error is only 16.1 yards for a 100 knot relative radial velocity. In the following discussion it will be assumed that ΔR is small enough so that it can be ignored.

C. NARROW BANDPASS IF FILTER

The optimum IF filter in terms of maximum signal-to-noise ratio is matched to the spectrum of the beat frequency in the IF amplifier. Since there is a different beat frequency for each possible target range, the optimum receiver must contain a large number of matched filters, one for each possible target range. For simplicity one IF filter can be used, and the spectrum of each target may then be shifted in sequence to that frequency by varying the frequency of the IF oscillator.

If the received echo signal is a pulse wave of finite duration, τ, and the pulse is repetitive and coherent, then the frequency spectrum of the wave has the shape shown in Figure 3 for the case where $T = 2\tau$. In practice the pulse train is finite, consisting of only one or a very few pulses, and successive pulses are not necessarily coherent. Moreover, there is a slow frequency modulation of the IF signal due to the sweep of the IF oscillator. The spectrum is therefore much more complex than shown in Figure 3 although its envelope is approximately in the SinX/X form shown.

In general, the beat frequency of the echo signal lies somewhere within a relatively wide band of frequencies. Let us assume that such a signal is brought to a bank of narrowband filters, as shown in Figure 5(a), whose combined range of frequencies covers the entire band of the

beat frequencies. Then one of the narrowband filters whose center frequency matches the frequency of that portion of the echo which is due to the target permits the identification and measurement of the target frequency and hence its range.

The bandwidth of an individual filter is wide enough to pass the signal, but not too wide to introduce more noise than need be. These filters intersect at their half-power points and have equal bandwidths as shown in Figure 5(b).

A true matched filter would have the SinX/X frequency response of Figure 3. It is not practical to build such a filter, so a compromise must be reached. A compromise that is often used is for the filter response to be similar to the central lobe of the SinX/X envelope.

II. RANGE SIDELOBE REDUCTION

The spectrum of a pulse signal has essentially SinX/X shape, with sidelobes extending on either side of the mainlobe. The first sidelobe is only 13.2 dB below the peak of the mainlobe, and the near sidelobes fall off at approximately 4 dB per sidelobe interval. Since these sidelobes may be mistaken for a target at a different range than that of the target, they are called range sidelobes. The range sidelobe null points are spaced $1/\tau$ hertz apart for the ideal SinX/X waveform. The range sidelobes of the strong target signals can obscure weaker signals as shown in Figure 6.

CW-FM radars exhibit in general a range ambiguity problem when signals are received from a large and a small target in the near vicinity of each other. In such cases the small target signals cannot be distinguished. An IF filter can respond to the range sidelobes of the strong-target signal spectrum. This spurious response could make it difficult to detect a small target such as a navigational buoy in the vicinity of a large object such as a ship.

The response of the receiver to the undesirable range sidelobes can be reduced by careful selection of the IF filter amplitude response. However optimizing the filter to minimize the range sidelobes will cause a reduction in output signal-to-noise ratio. The optimum filter for

maximum output signal-to-noise ratio would be matched to the signal spectrum. Increasing the output signal-to-noise ratio by optimum filtering improves the ability of the receiver to detect the presence of a weak signal. In order to make the target echo signal relatively more detectable, a narrow bandpass filter centered at frequency $f_{IF} + f_b$ should be employed to pass the signal which falls within its passband but rejects all noise outside its passband. The maximum peak output signal-to-noise ratio in terms of the signal energy for the optimum matched filter is given in Refs. 2 and 3.

$$\left(\frac{S}{N}\right)_{out} = \frac{2E}{N_o}$$

E is the signal energy, N_o is the noise power density in terms of watts per hertz.

The range sidelobes may be reduced either by amplitude weighting the envelope of the transmitted FM signal or by an amplitude weighting of the receiver filter frequency response. Proper shaping of the envelope of the signals can significantly reduce the sidelobe levels. Figures 7 and 8 [Ref. 4] show that a reduction in sidelobe levels is accomplished by employing the truncated Gaussian and truncated Butterworth signals in an FM chirp (low duty cycle) pulse compression radar. For the amplitude weighting shown range sidelobes are down 35 dB, whereas without weighting they

would have been down only 13 dB. Figure 9 [Ref. 3] shows us typical chirp radar receiver output waveforms resulting from use of the frequency weighting techniques in which the response far from the carrier falls off more rapidly than does matched filter response. These waveforms are obtained from the weighting function

$$W(\omega) = k + (1 - k)\mathrm{Cos}^n \frac{\pi(\omega - \omega_o)}{\Delta\omega}$$

Because of its effect on the shape of this frequency response, k is called the pedestal function. The weighting function has two effects on the system output. First, the amplitude of range sidelobes is reduced by an amount depending on the values of k and n. Second, since the combined filter is no longer a matched filter, the signal-to-noise ratio is slightly reduced and the pulse width increased. Increased width and the sidelobe reduction are shown in Figure 9 for several values of n and k.

In the present study an attempt is made to optimize the relative widths of the spectrum and the filter to reduce the amplitude of the range sidelobes in the case of the FM-CW radar. Here, the IF spectrum is quite different than for the FM-CW radar. The question is, can range sidelobes be reduced in this case by the same or different techniques?

III. EXPERIMENTAL RESULTS

In order to determine the effect of the relationship between filter bandwidth and the IF pulse width and duty cycle, the experimental set-up shown in Figure 10 was used. The frequency synthesizer generated a CW signal similar to the signal in the IF amplifier of the FM-CW radar. The pulse modulated IF signal shown in Figure 11 was fed to the input of the filter. The output of the filter was measured by means of the power meter. The time response of the filter was displayed on an oscilloscope as shown in Figure 12. By changing the width and duty cycle of the input signal pulse various filter-output responses were obtained. The responses obtained by changing the IF frequency in 100 hertz steps to simulate different target ranges are shown in Figures 16 through 30. Evidently, the pulse width and duty cycle affect the response of the filter and the amplitude of the range sidelobes in the filter output.

A careful examination of the IF filter responses shown in Figures 16 through 30 shows that both the width and the magnitude of the range sidelobes of the IF filter response are considerably affected by the period and duration of the pulse modulated IF signal. It is also seen that for the conditions of Figure 20 the first order range sidelobes are lowest with respect to the peak. In this case the

signal spectrum shape almost matches the filter spectrum shape. In this particular case (mention parameters-pulse width, duty cycle), the half-amplitude bandwidth of the filter happens to almost coincide with the bandwidth of the main lobe of the signal. It is also noted from Figure 20 that the first range sidelobe of the filter output response is 11.1 dB lower than the main lobe and lower than the sidelobes obtained in all other figures. This reduction has been achieved, as noted earlier, by matching the shape of the received wave form spectrum to the filter spectrum shape. Thus, the maximum energy is transferred from the filter when the shape of the central portion of the signal spectrum matches the filter spectrum shape. As a by-product of this matching, the relative amplitude of the range sidelobes is reduced.

It is also seen from Figures 16 through 30 that the width of the response, δ_f, is narrowest for the Figure 20 case. Since

$$f_b = \frac{2R}{c} \cdot \frac{\Delta f}{T}$$

it follows that

$$\delta_{f_b} = \delta_f = \frac{2\Delta R}{c} \cdot \frac{\Delta f}{T}$$

Solving for ΔR, we have an expression for range resolution

$$\Delta R = \frac{cT}{2\Delta f} \delta_f$$

Thus the conditions of Figure 20, spectrum matched to filter width, also give the best range resolution.

IV. ANALYTICAL SOLUTION

A. COMPUTATION OF POWER SPECTRA

1. The Fourier Transform

Suppose that a periodic time function f(t), having duration τ, is to be represented by a Fourier series. The form of the series is then given by

$$f(t) = \frac{a_o}{2} + \Sigma[a_n \mathrm{Cos}(2\pi nft) + b_n \mathrm{Sin}(2\pi nft)] \tag{1}$$

in which the coefficients a_n and b_n are

$$a_n = \frac{2}{T} \int_o^T f(t)\ \mathrm{Cos}(2\pi nft)dt \tag{2}$$

$$b_n = \frac{2}{T} \int_o^T f(t)\ \mathrm{Sin}(2\pi nft)dt \tag{3}$$

and f = 1/T is the fundamental frequency. Non-periodic waves can be analyzed through the use of the Fourier integral which is an extension of the Fourier series. Such a function may be transformed from the time domain to the frequency domain and vice versa by the following transform pair.

$$f(t) = \frac{1}{2\pi} \int_{-\infty}^{+\infty} F(f)\ e^{j2\pi nft}\ df \tag{4}$$

$$F(f) = \int_{-\infty}^{+\infty} f(t)\, e^{-j2\pi nft}\, dt \tag{5}$$

The function F(f) is called the Fourier transform of f(t), and f(t) is called the inverse Fourier transform of F(f). The integral in F(f) analyzes the time function into a continuous spectrum.

Over a time $0 \leq t \leq \tau$, the power spectrum of f(t) will be estimated by squaring the absolute value of the Fourier transform F(f). No signals have been taken outside the interval $0 \leq t \leq \tau$ so that the integration needs to be performed only over this time interval.

The spectra can be determined by using the Fourier transform formula for F(f). Then the Fourier transform is computed and the spectral power density is obtained as $F^2(f)$. Let us illustrate this procedure by applying it to a sinusoidal pulse of length τ. Defining f(t) as

$$f(t) = \begin{cases} A\,\mathrm{Sin}(2\pi f_o t) & -\tau/2 \leq t \leq \tau/2 \\ 0 & |t| > \frac{\tau}{2} \end{cases} \tag{6}$$

and inserting Eq. (6) into Eq. (5) one has

$$F(f) \equiv A \int_{-\tau/2}^{\tau/2} \text{Sin } 2\pi f_o t \; e^{-j2\pi ft} \, dt \tag{7}$$

Writing $\text{Sin} 2\pi f_o t$ as,

$$\text{Sin } 2\pi f_o t = \frac{e^{j2\pi f_o t} - e^{-j2\pi f_o t}}{2j} \tag{8}$$

the integral in Eq. (7) takes the form,

$$F(f) = \frac{A}{2j} \int_{-1/2}^{\tau/2} [e^{2\pi(f_o - f)t} - e^{-2\pi(f_o - f)t}] \, dt \tag{9}$$

After integration, $|F(f)|^2$ becomes

$$|F(f)|^2 = [\frac{A^2\tau^2}{4}] [\frac{\text{Sin}\pi(f_o - f)\tau}{\pi(f_o - f)\tau} + \frac{\text{Sin}\pi(f_o + f)\tau}{\pi(f_o + f)\tau}]^2 \tag{10}$$

A part of the spectral power density given by Eq. (10) is shown in Figure 4. There is an identical spectrum for negative frequencies.

2. <u>Power Transform through a Network</u>

Suppose that a periodic signal f(t) is applied to a filter with a transfer function H(f) and the output $f_o(t)$ obtained from the product of the Fourier series and the transfer function

$$f_o(t) = \sum_{n=-\infty}^{\infty} H(nf_o) F\, e^{j2\pi n f_o t} \tag{11}$$

The power of f(t) is given by (assuming R = 1 ohm)

$$S = \sum_{n=-\infty}^{\infty} |F(f)|^2 \tag{12}$$

while the power of $f_o(t)$, the filter output reduces to

$$S_o = \sum_{n=-\infty}^{n=\infty} |H(nf_o)|^2 \; |F(f)|^2 \tag{13}$$

B. ANALYTICAL SOLUTION

For the analytical solution an analytic filter characteristic was selected with a 6 dB bandwidth about the same as the experimental filter characteristic.

Figures 13 and 14 show us the characteristics of the physical and analytic filters, respectively. The transfer function of the two-pole filter is given by Eq. (14),

$$H(f) = \frac{J\omega KQG_m}{4\left[\left(\frac{f-f_o}{f}\right)Q\right]^2 - 1 - K + j4\left(\frac{f-f_o}{f}\right)Q} \tag{14}$$

Here K is the smoothing factor of the filter characteristic. K was taken to be unity, Gm is a constant scaled to make the maximum gain unity, and $f_o = 30$ MHz.

The magnitude of the output of the filter was next computed for various frequency offsets, in 200 hertz steps between the center of the filter. This is equivalent to moving the filter across the spectrum as was done experimentally. Convolution was employed for this computation, and the computation was made for four different combinations of pulse width and period. Figure 15 shows the computed responses in each case. It is also seen from Figure 15 that range sidelobe amplitude levels are reduced by matching the mainlobe of the signal to the filter spectrum shape.

V. CONCLUSIONS

It has been shown that the amplitudes of the range sidelobes in the output of the FM-CW radar of Figure 2 are a function of the relative widths of the main lobe of the IF spectrum and of the width of the filter passband. Experimental work employed a four-pole filter while theoretical work used a two-pole filter. In both cases the range sidelobe amplitude proved to be minimal when the 6 dB filter width matched the 6 dB width of the central lobe of the signal spectrum. With the four-pole filter used experimentally the total signal fell off more rapidly with change of range than two-pole filter.

It is probable that range sidelobes can also be reduced by amplitude weighting of the signal. Time did not permit an investigation of this.

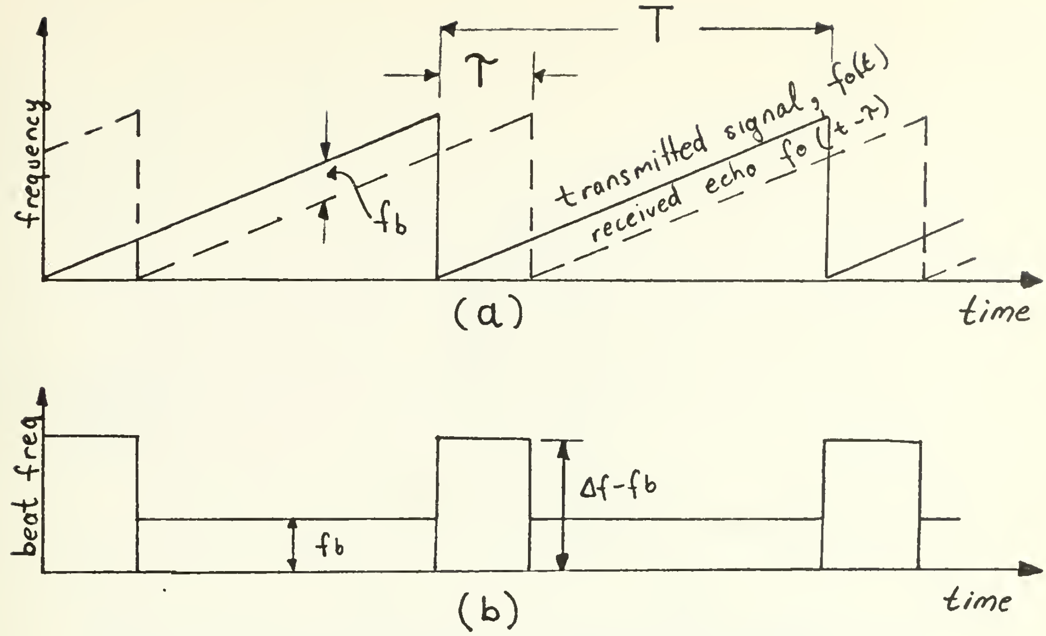


Figure 1 (a) Sawtooth frequency modulation
(b) Beat note

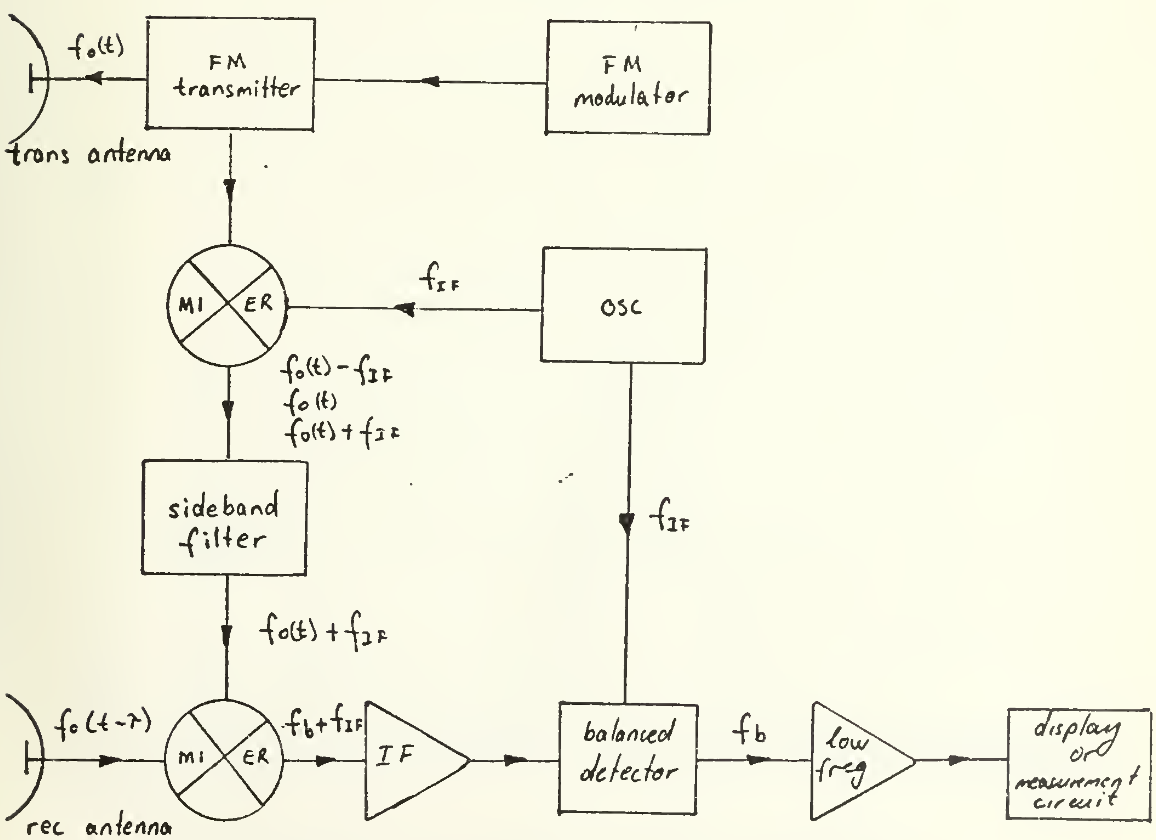


Figure 2 Block diagram of CW-FM radar

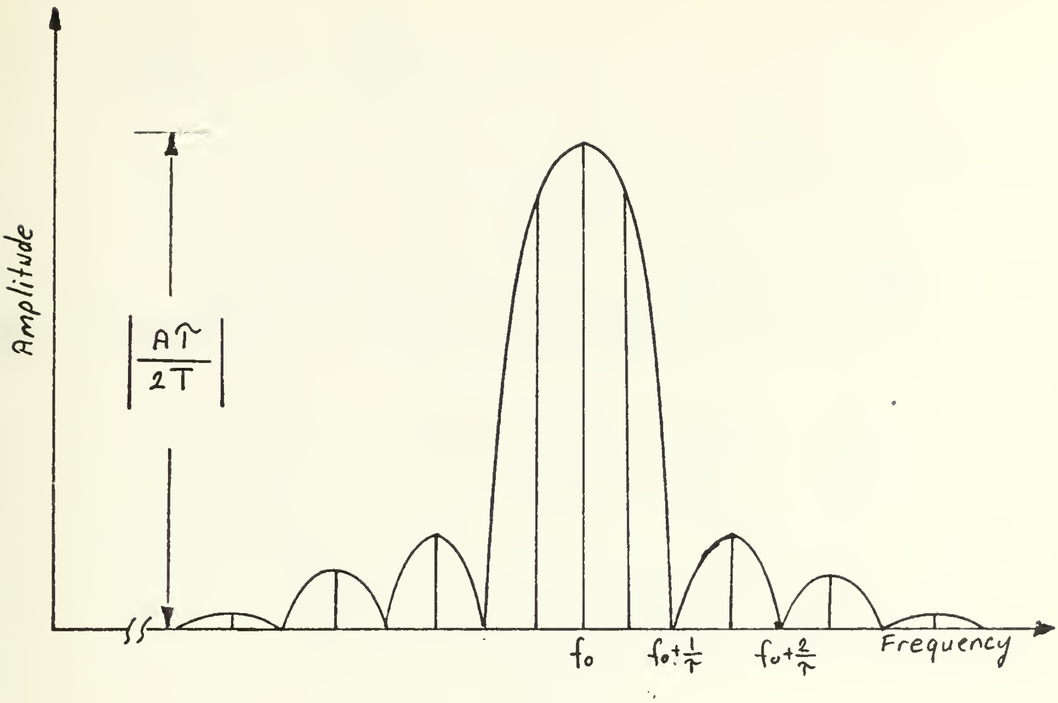


Figure 3 Frequency spectrum of a repetitive square wave pulse

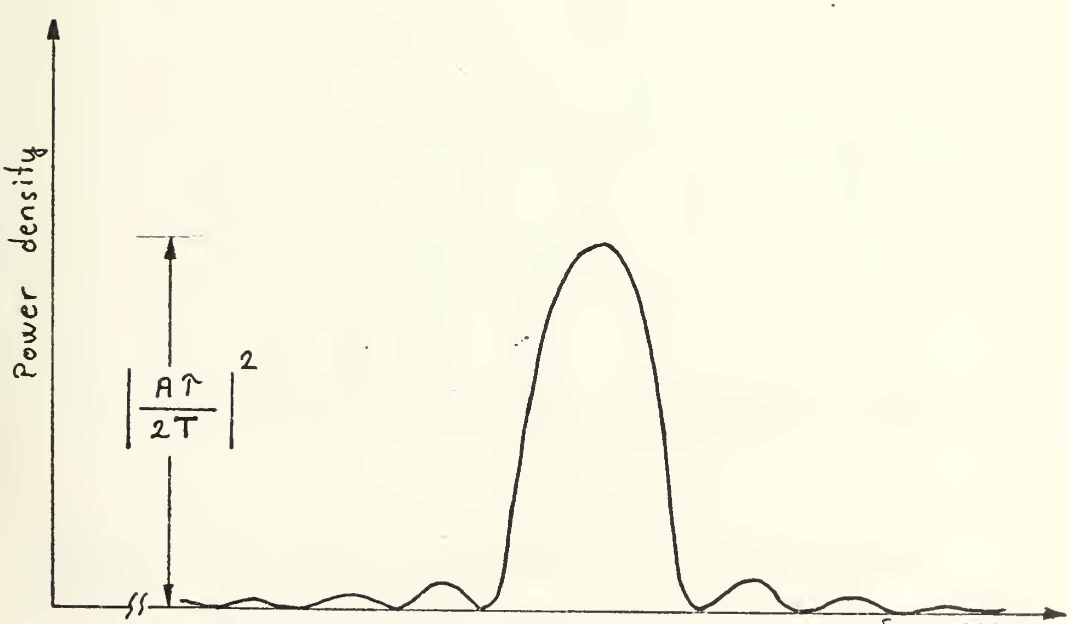
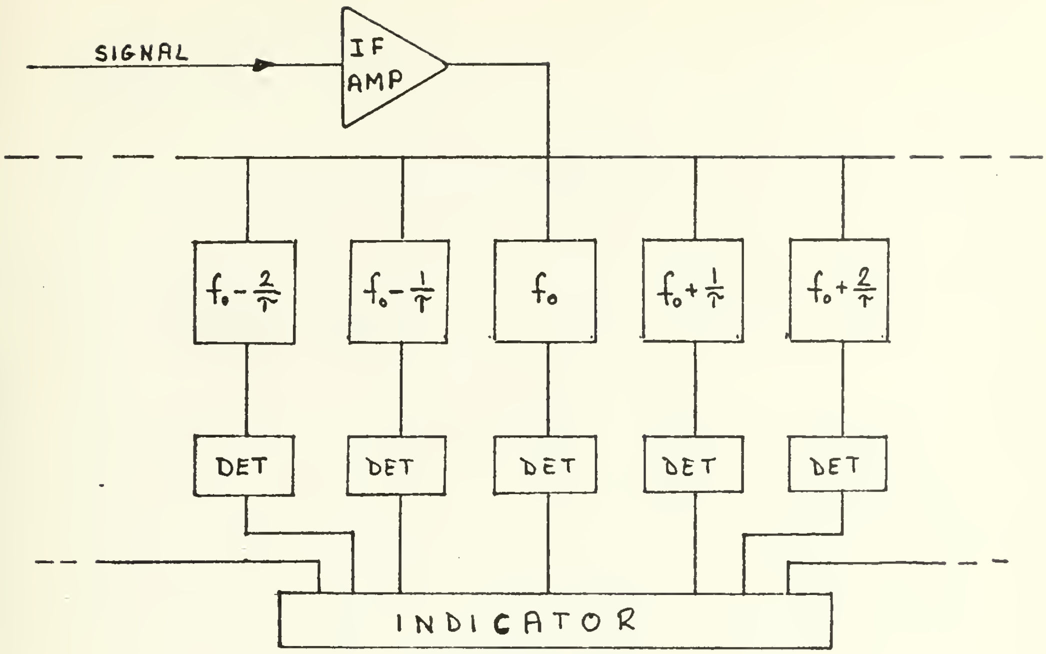
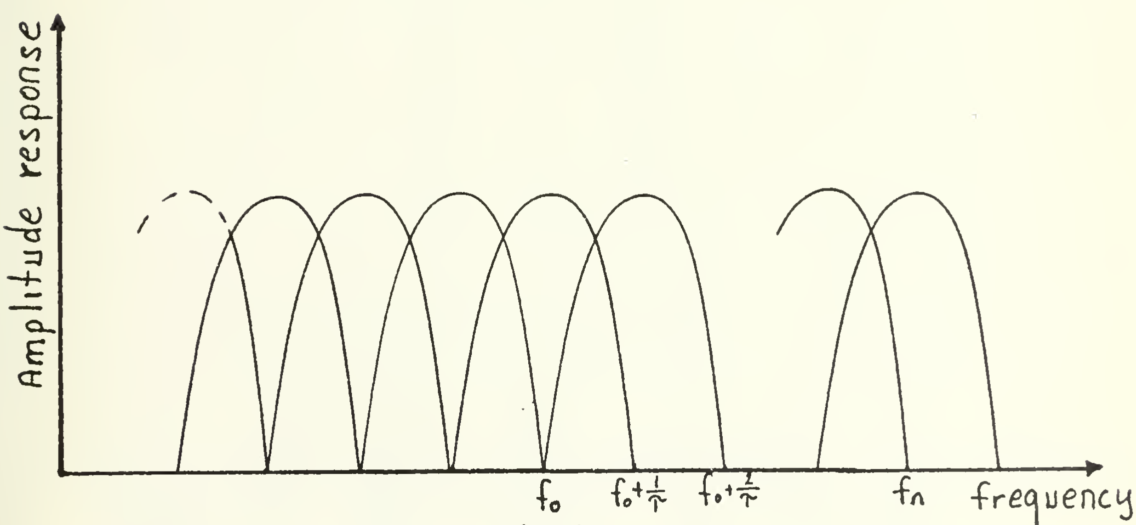


Figure 4 Power spectral density



(a)



(b)

Figure 5 (a) Block diagram of IF filter bank
(b) Frequency response of IF filter bank

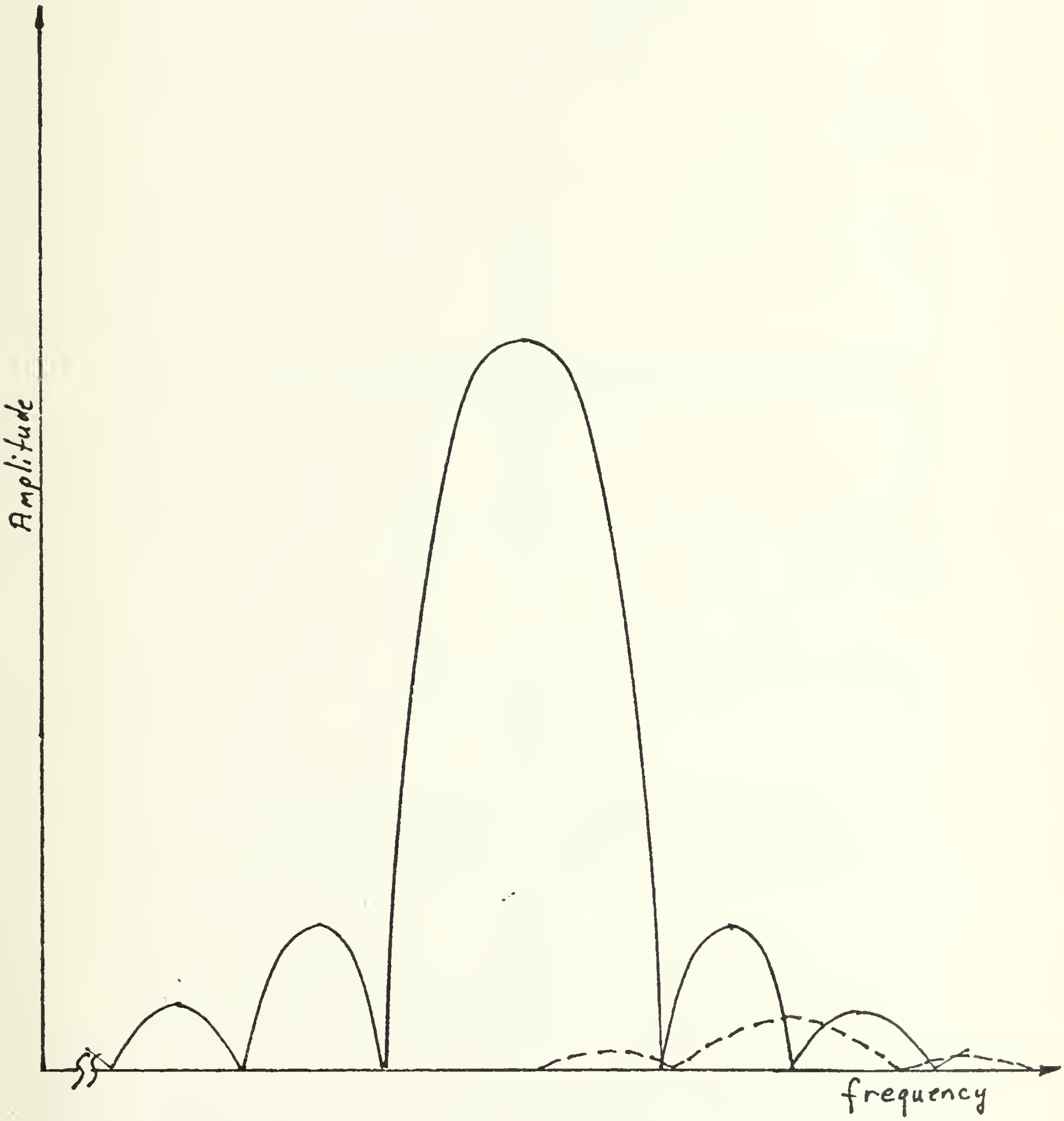


Figure 6 Effect of range sidelobes on weak signal detection

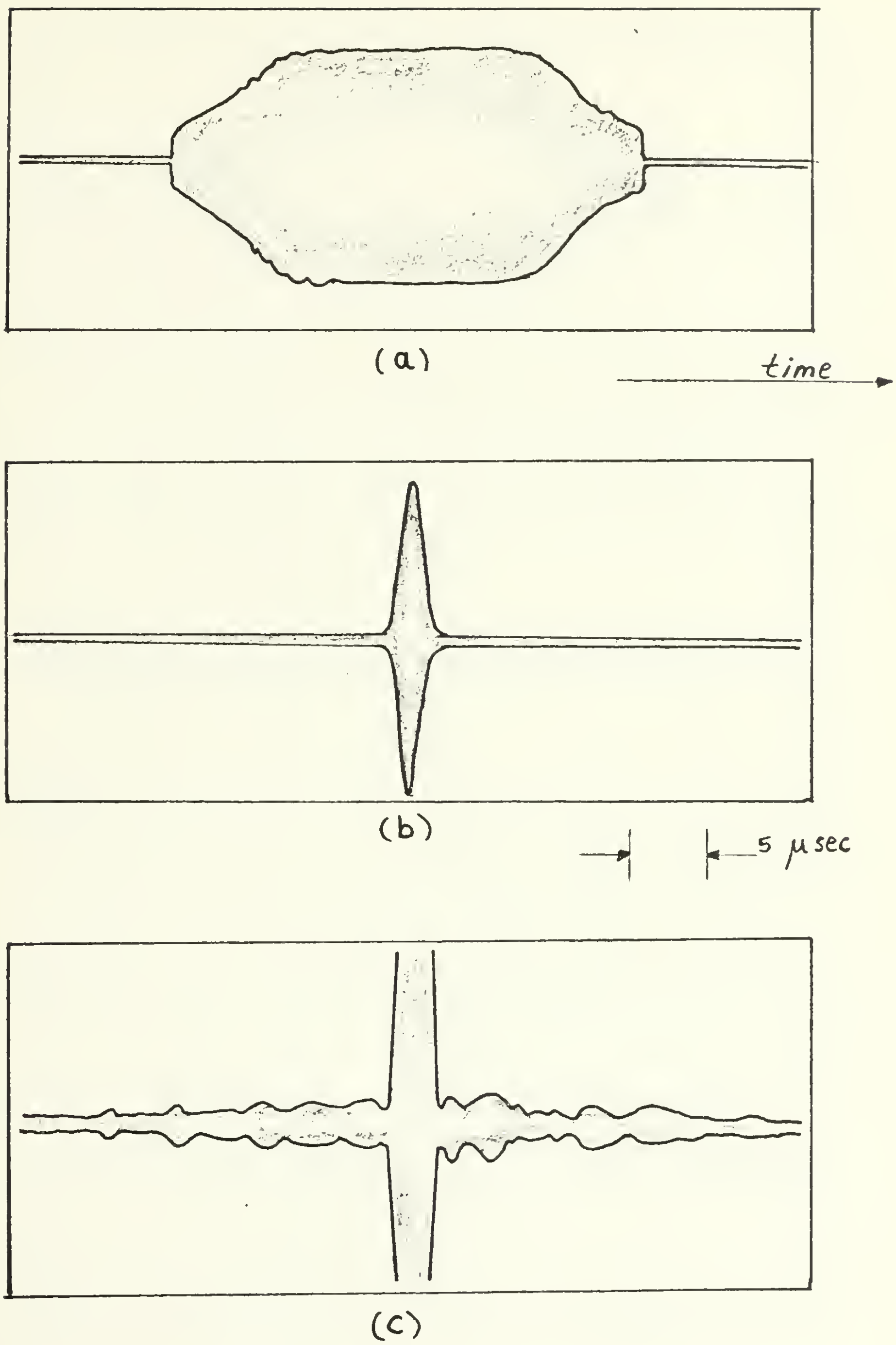


Figure 7 The receiver output waveform the truncated Butterworth case

(a) Butterworth signal

(b) Receiver output waveform

(c) Receiver output waveform amplified 20 times

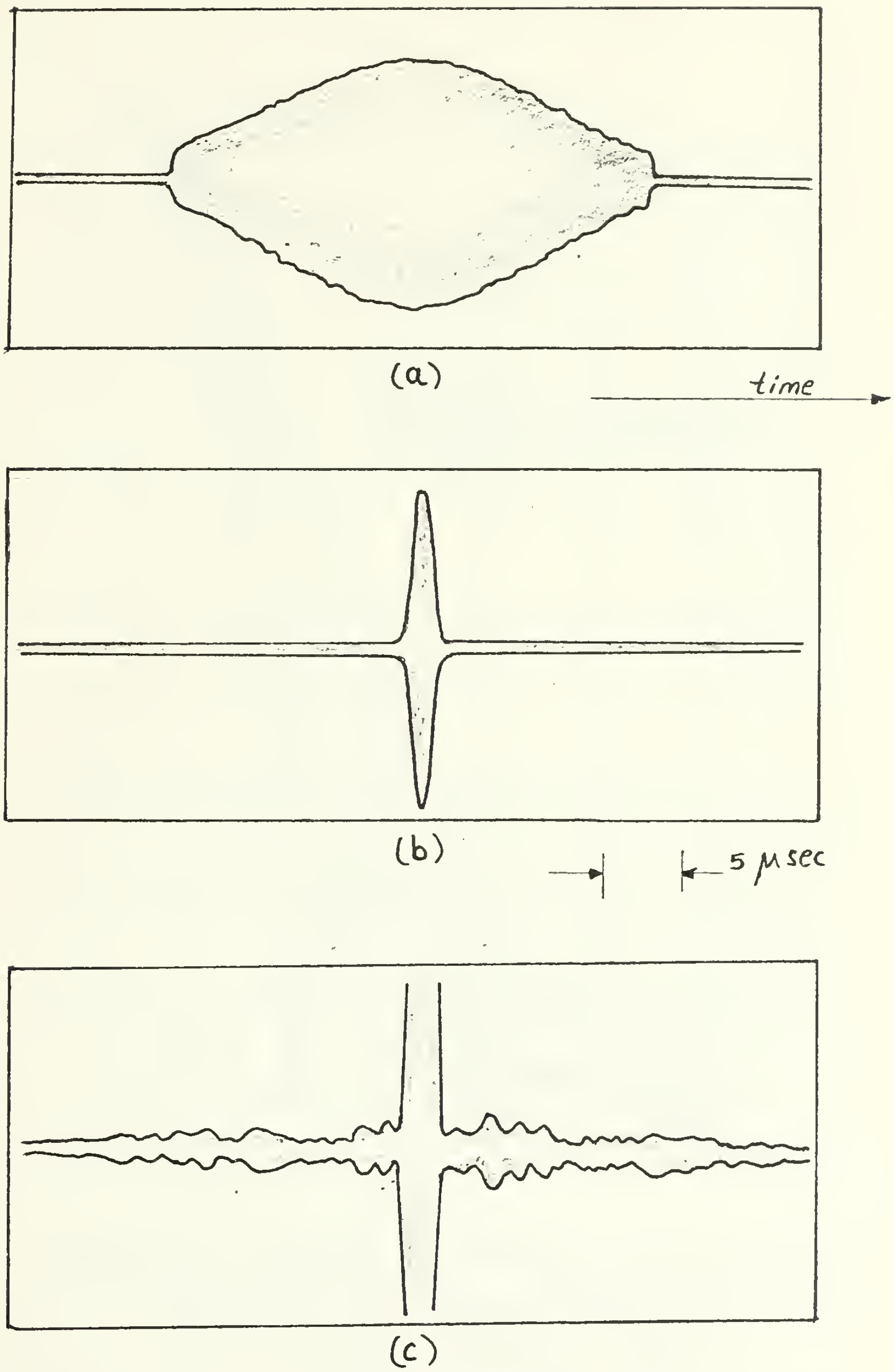


Figure 8 The receiver output waveform for the truncated Gaussian case

(a) Gaussian signal

(b) Receiver output waveform

(c) Receiver output waveform amplified 20 times

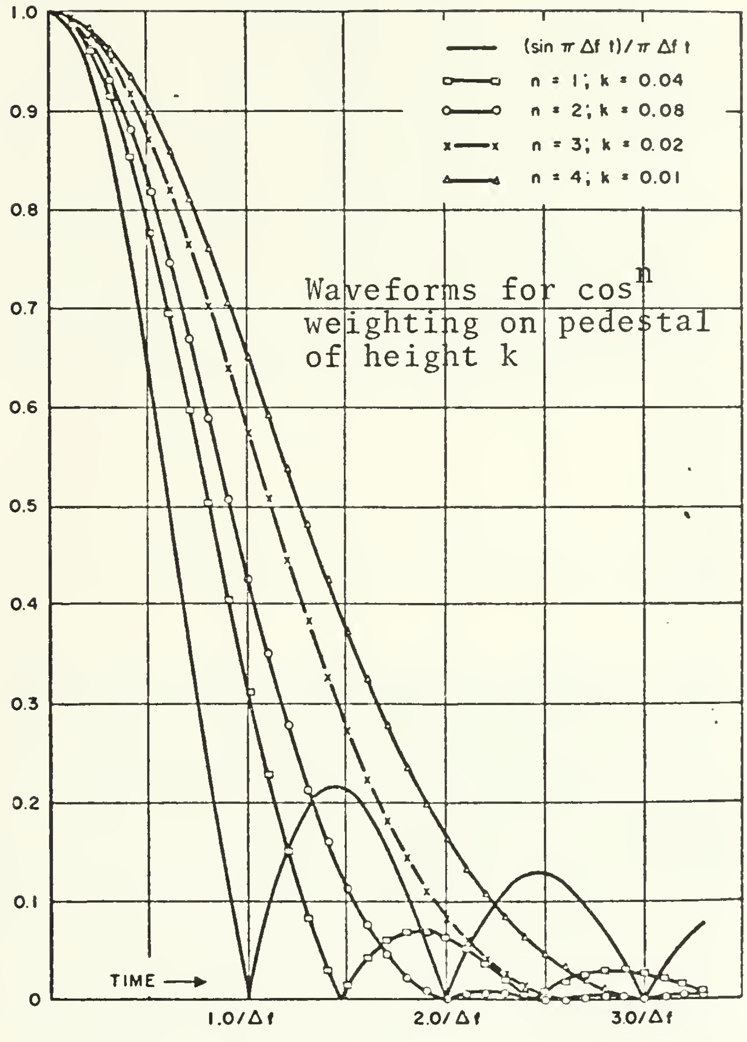


Figure 9 Typical waveforms resulting from frequency weighting techniques at the matched-filter output [Refs. 2, 3]

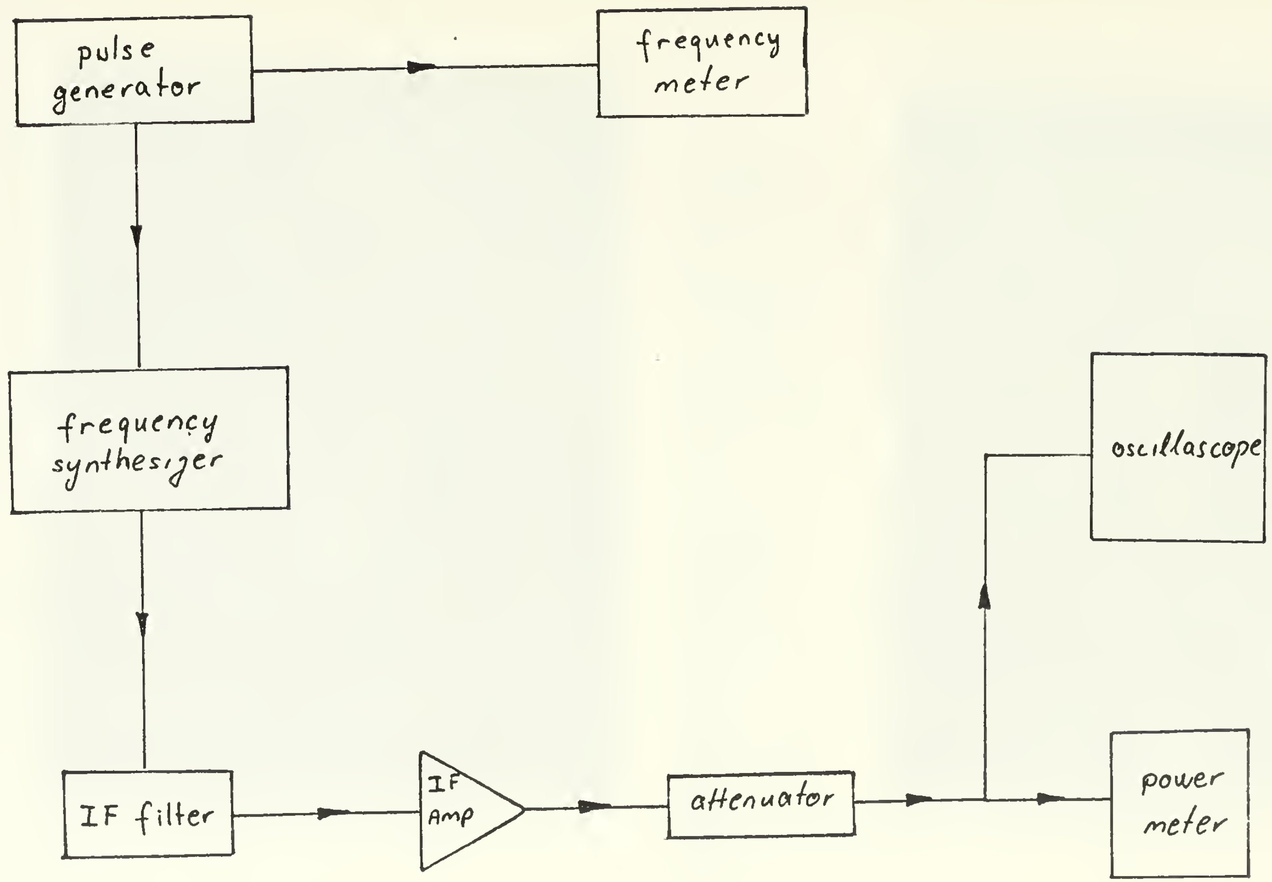


Figure 10 Block diagram of the experimental setup

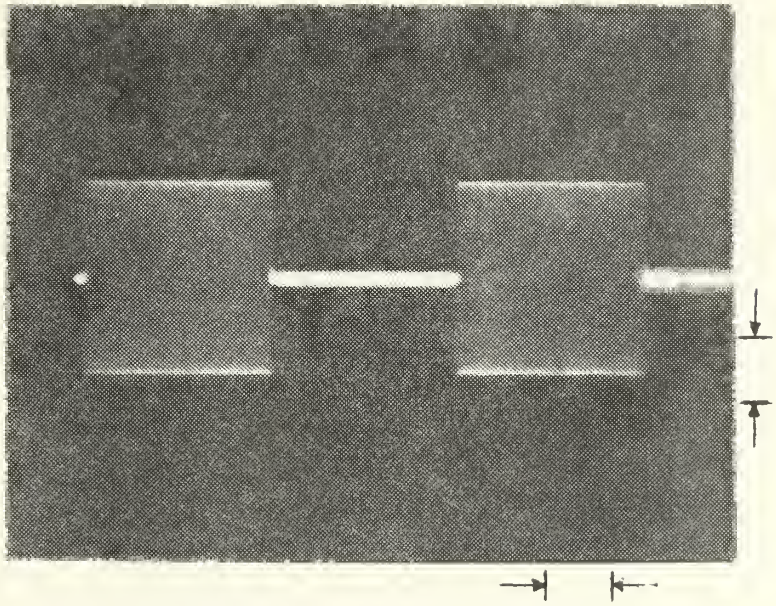


Figure 11 Pulse modulated IF carrier
Vertical scale : 1 volt per division
Horizontal scale: 100 μsec per division

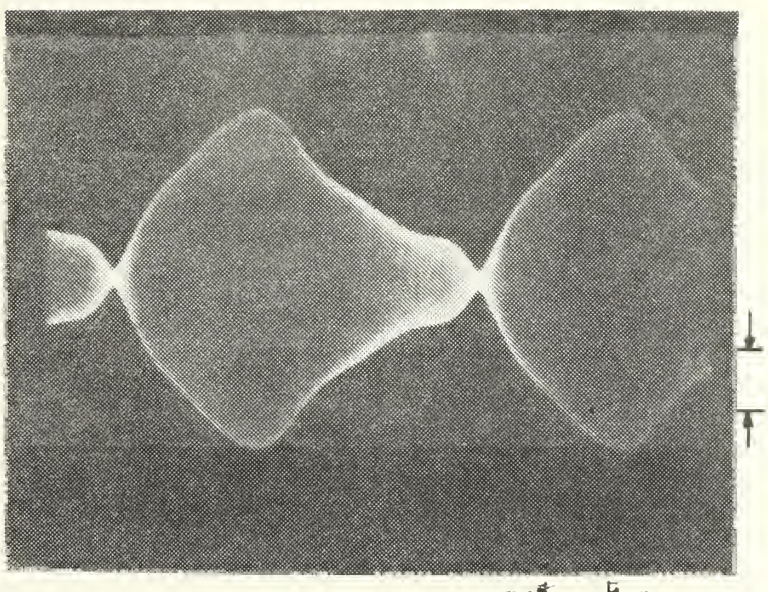


Figure 12 IF filter response
Vertical scale : 0,01 volts per division
Horizontal scale: 200 μsec per division

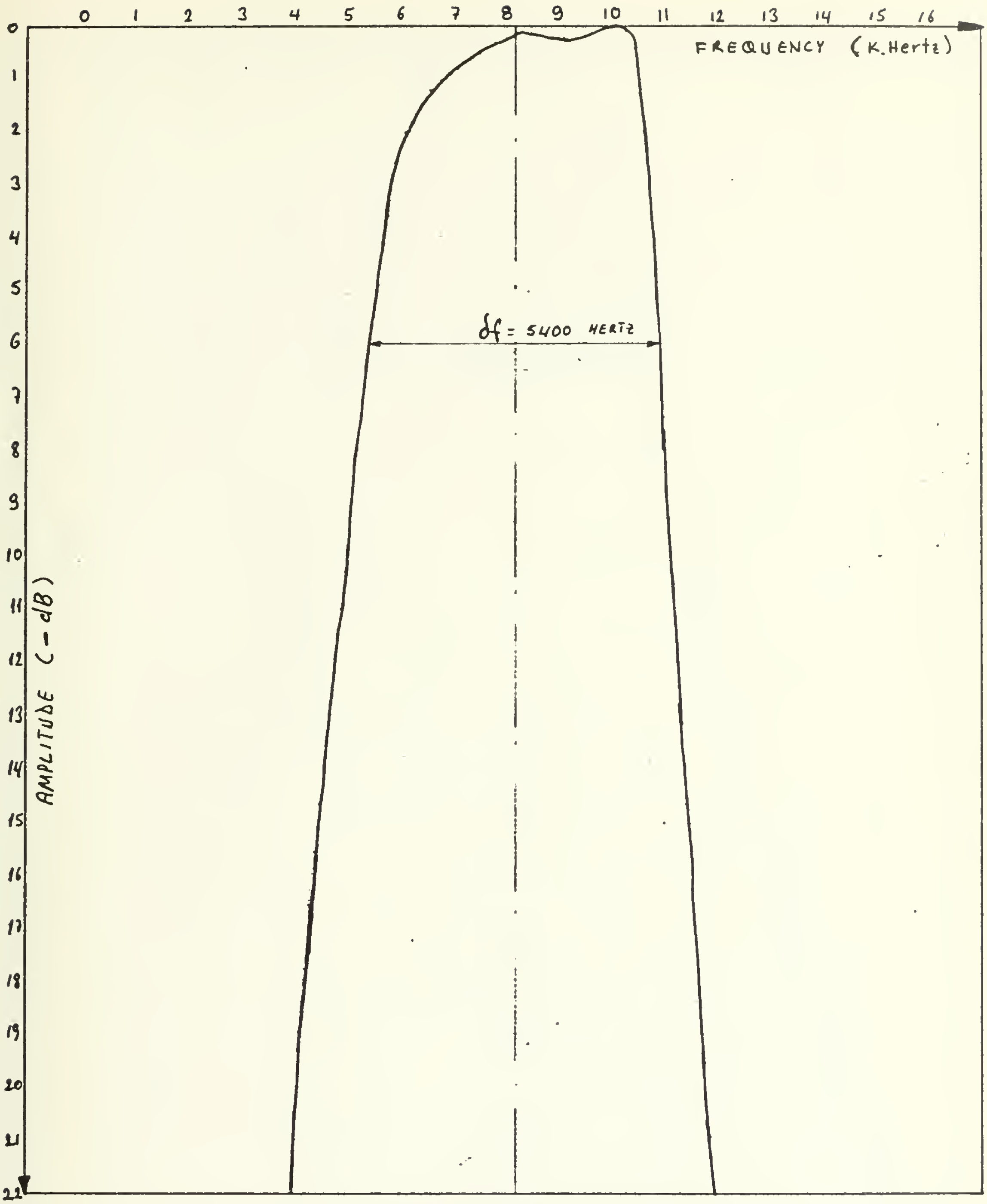


Figure 13 High-Q quartz crystal narrow bandpass IF filter response

Center frequency: 29,998,200 Hertz

δf: 5400 Hertz

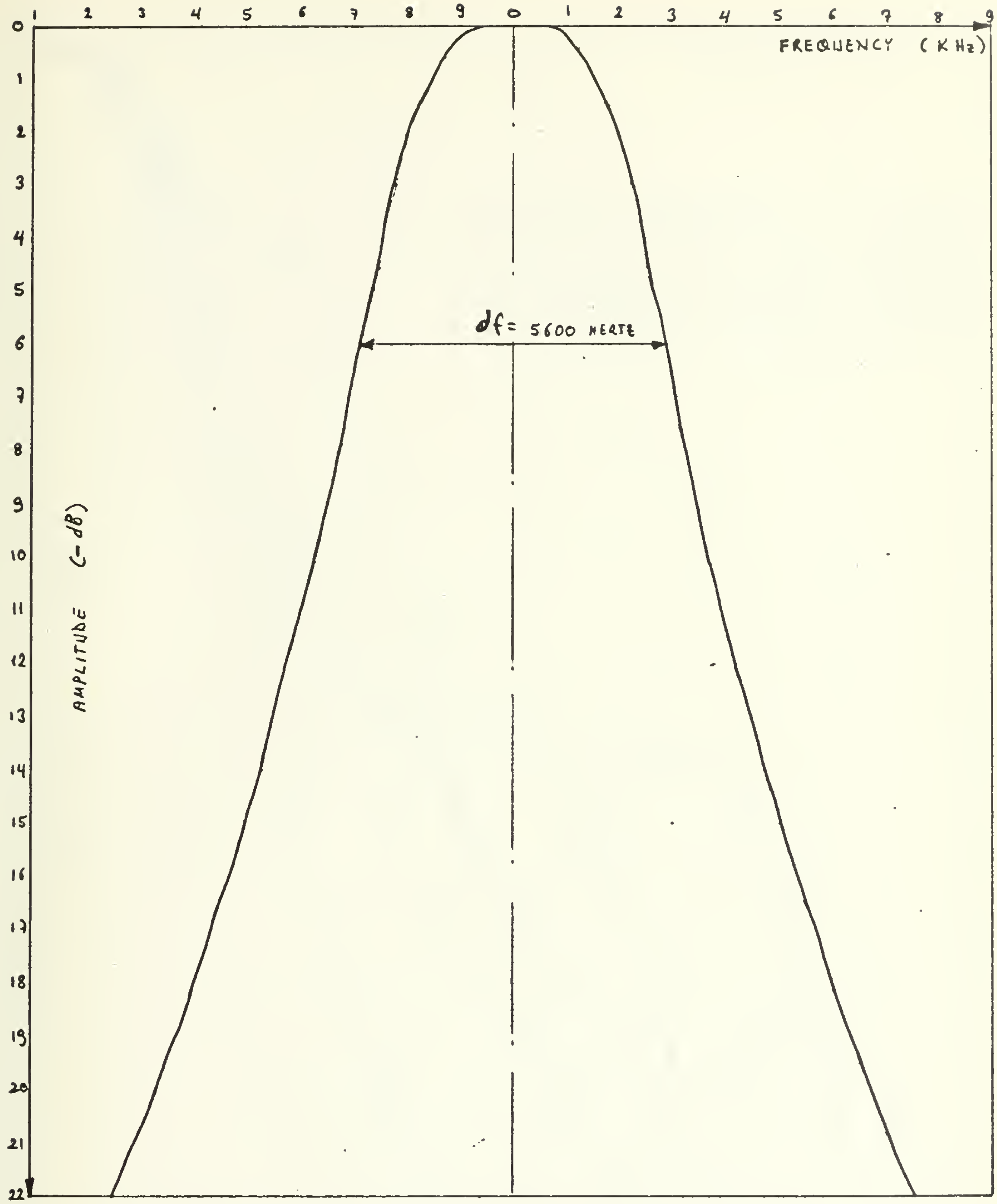


Figure 14 Analytical IF filter characteristic

Center frequency: 30,000,000 Hertz

δf: 5600 Hertz

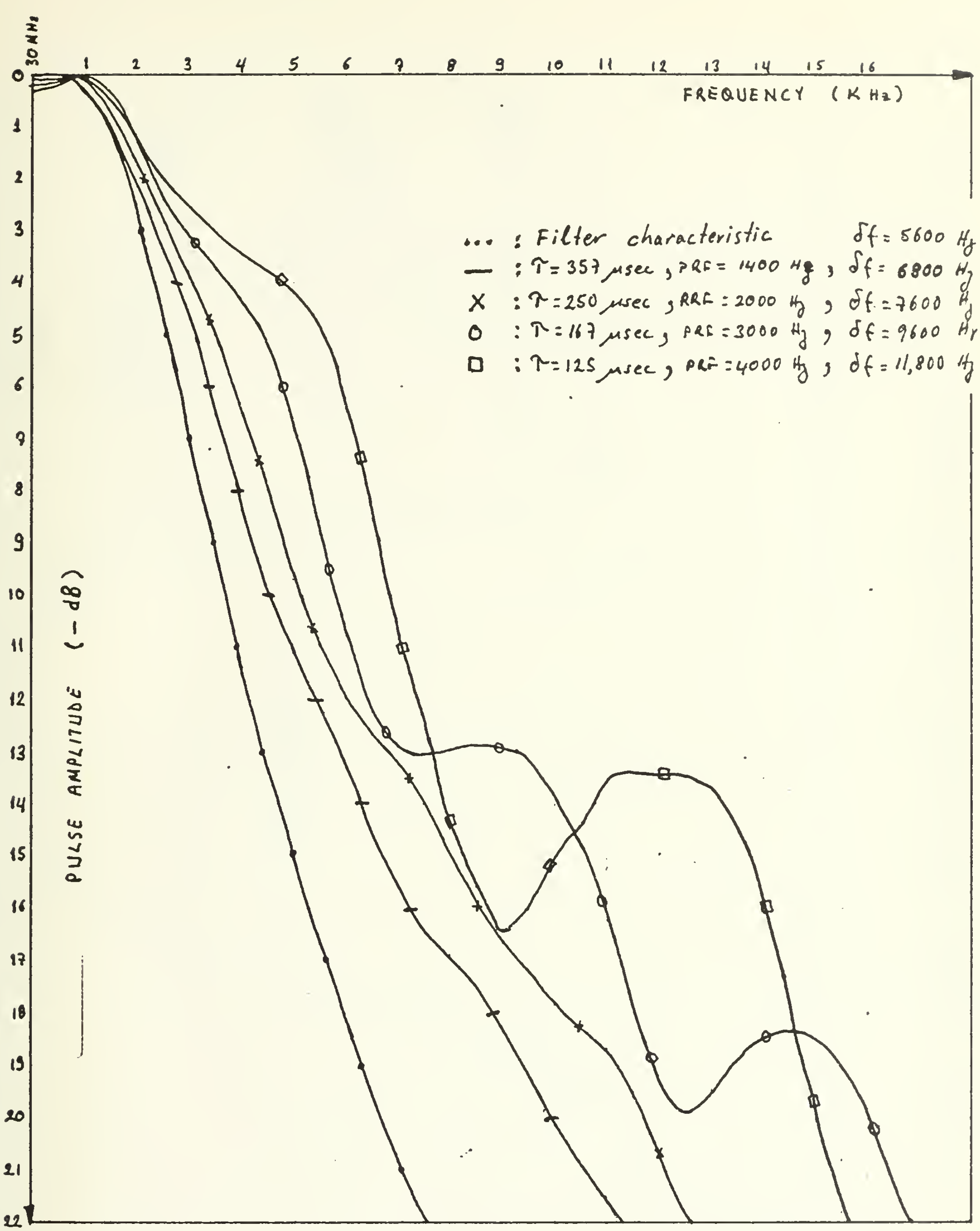


Figure 15 Effects of the pulsewidth of the pulse signal on the IF filter

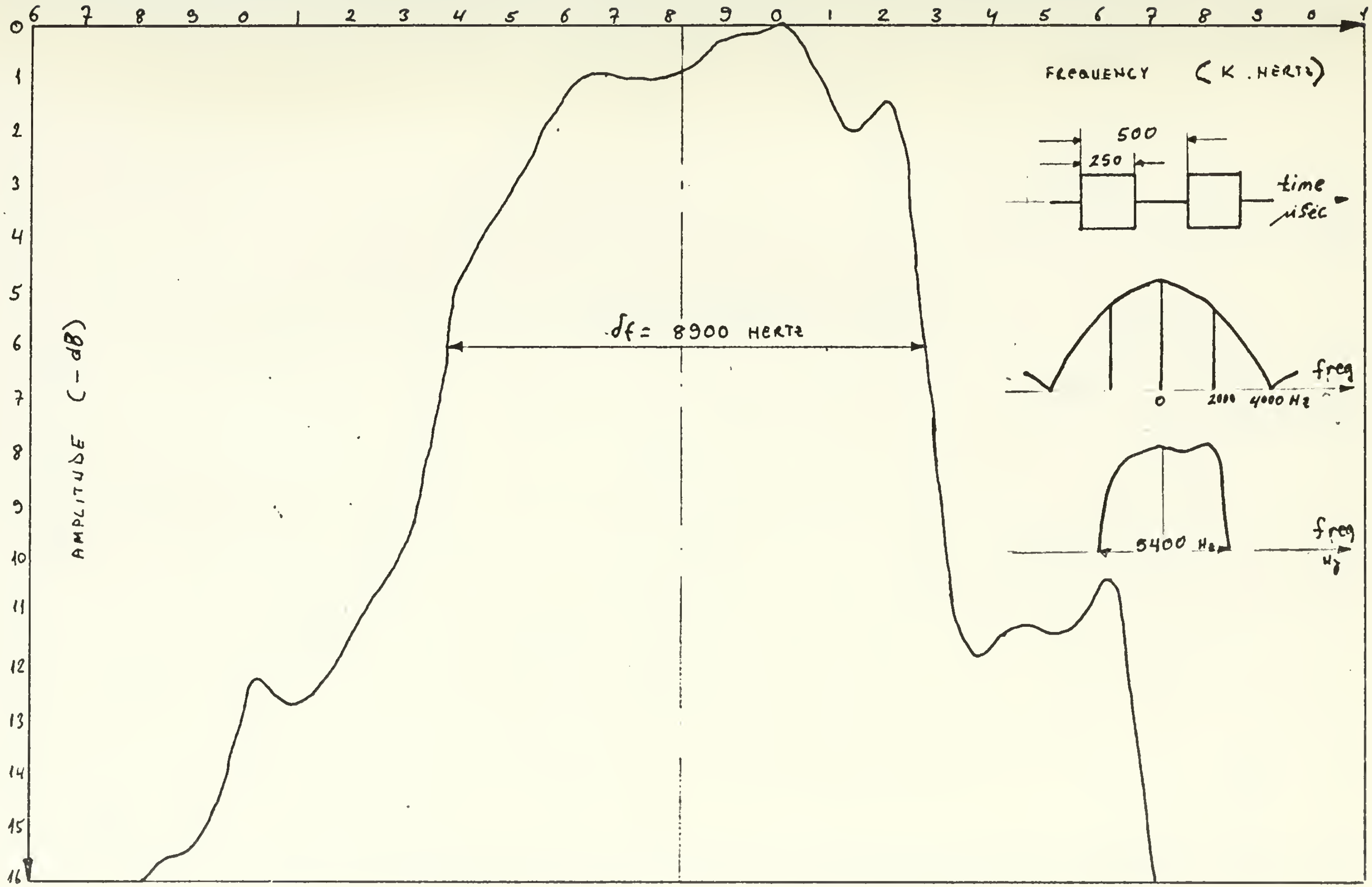


Figure 16 The effect of the width of the input pulse signal on the IF filter
First sidelobe 10.3 dB lower than the mainlobe
Input pulse width τ = 250 μsec
Duty cycle = 1/2

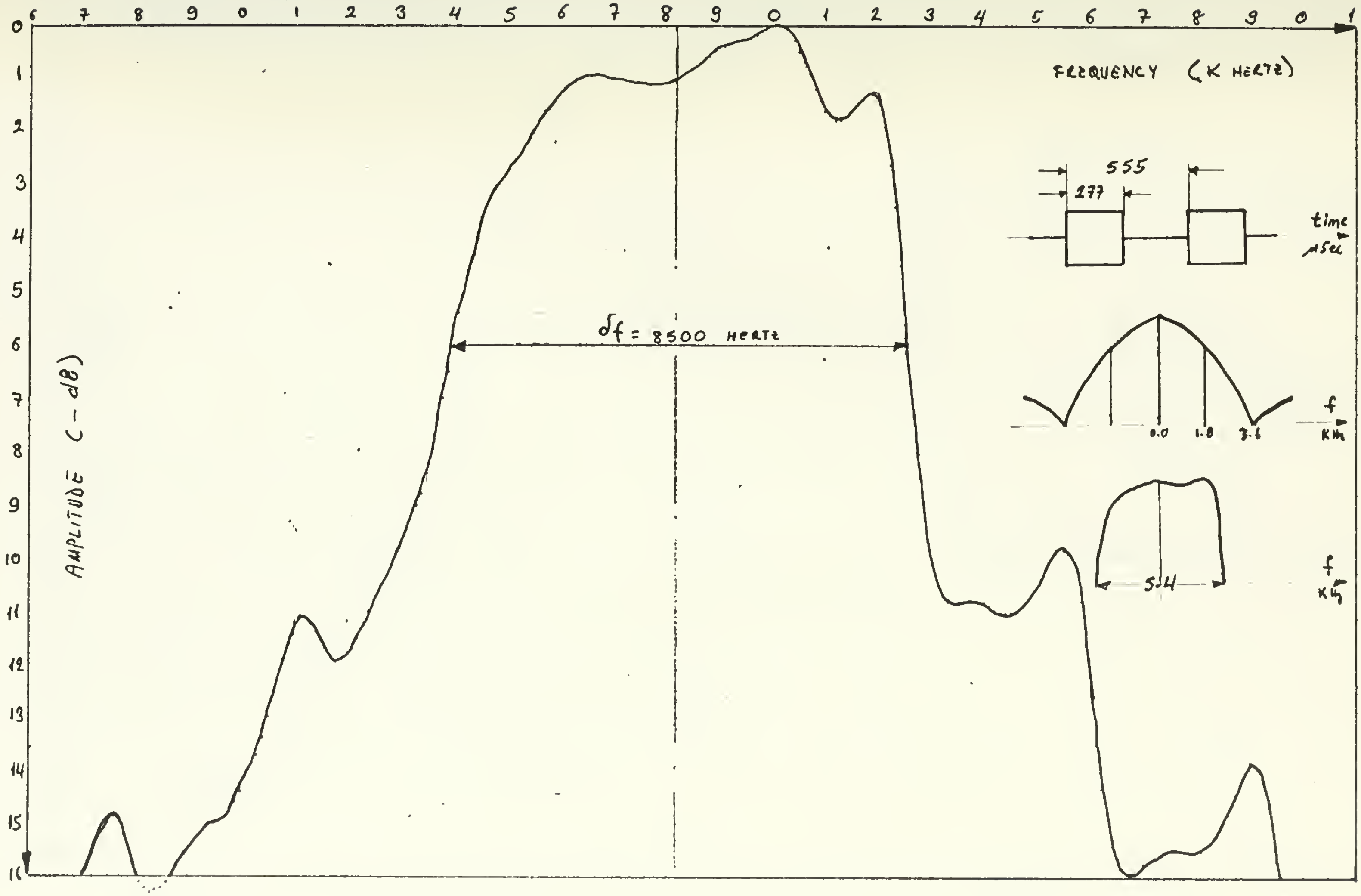


Figure 17 The effect of the width of the input pulse signal on the IF filter
First sidelobe 9.8 dB lower than the mainlobe
Input pulse width τ = 277 µsec
Duty cycle = 1/2

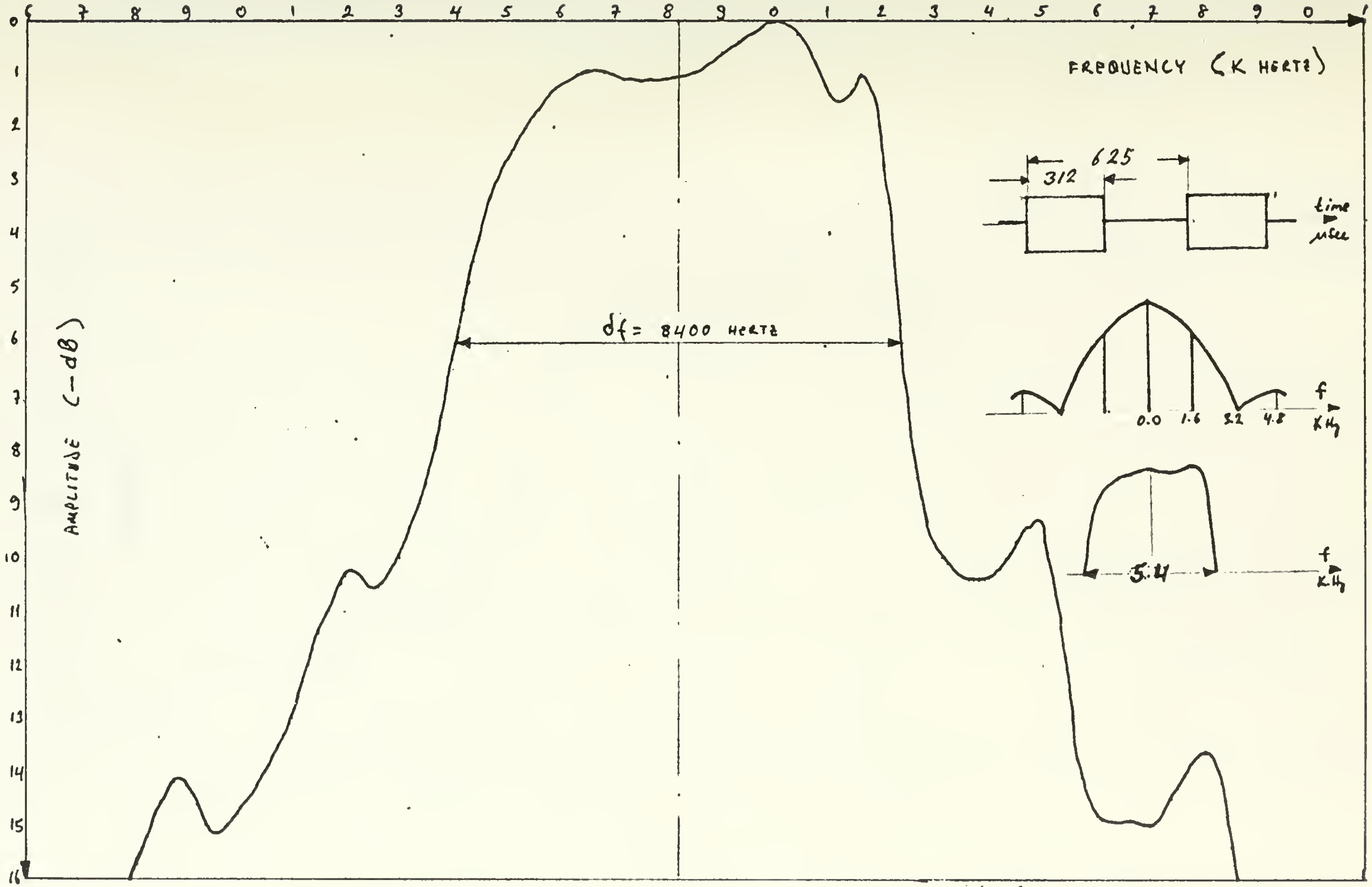


Figure 18 (a) The effect of the widt- of the input pulse signal on the IF filter
First sidelobe 9.3 dB lower than the mainlobe
Input pulse width τ = 312 µsec
Duty cycle = 1/2

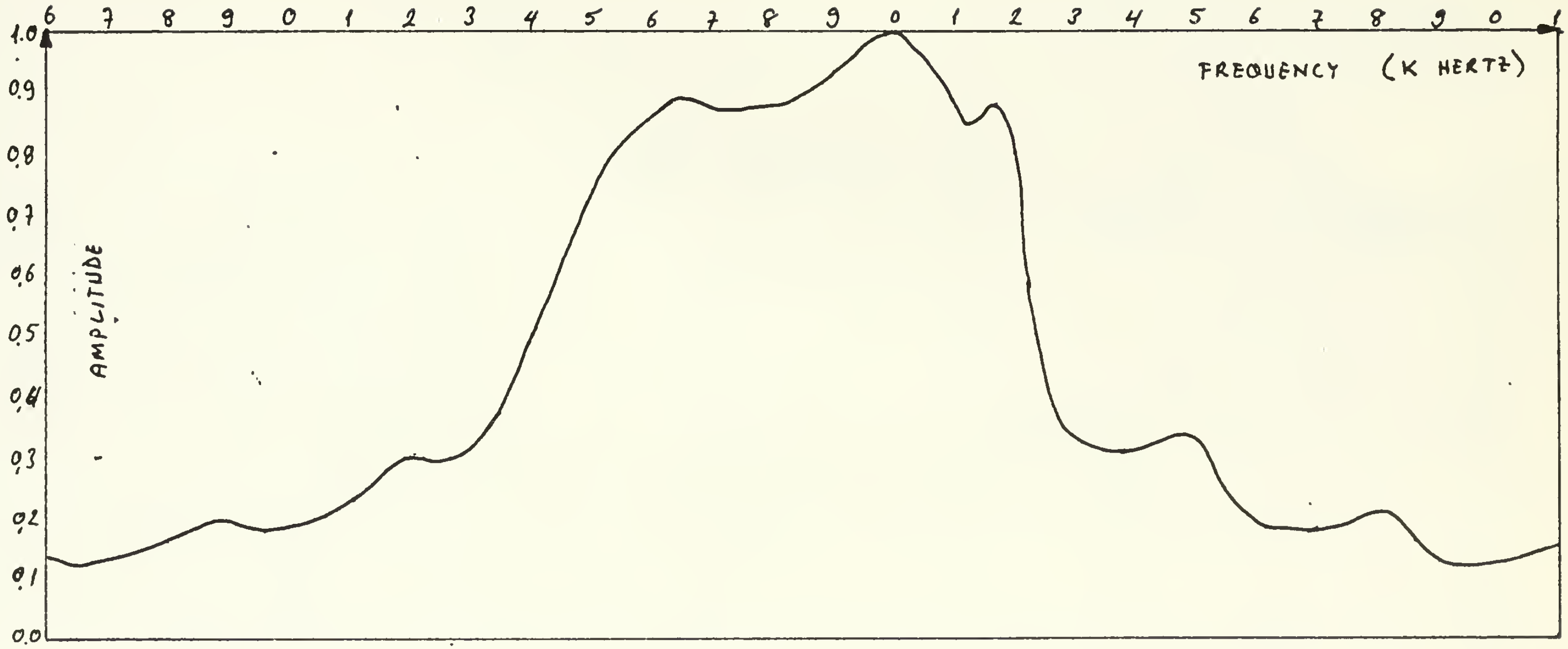


Figure 18 (b) The effect of the width of the input pulse signal on the IF filter Figure 18(b) shows the same information on a linear scale for comparison.

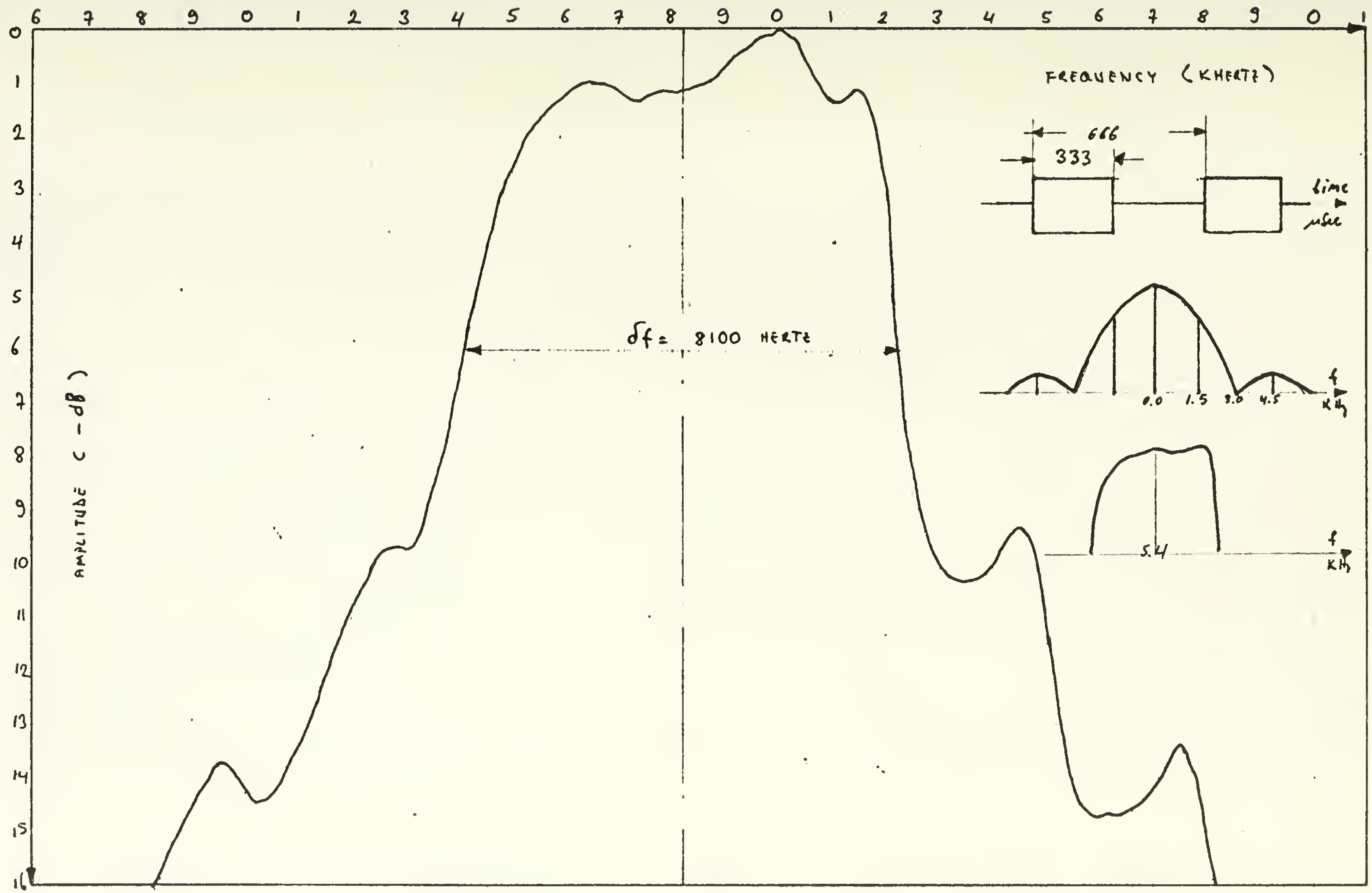


Figure 19 (a) The effect of the width of the input pulse signal on the IF filter
First sidelobe 9.3 dB lower than the mainlobe
Input pulse width τ = 333 μsec
Duty cycle = 1/2

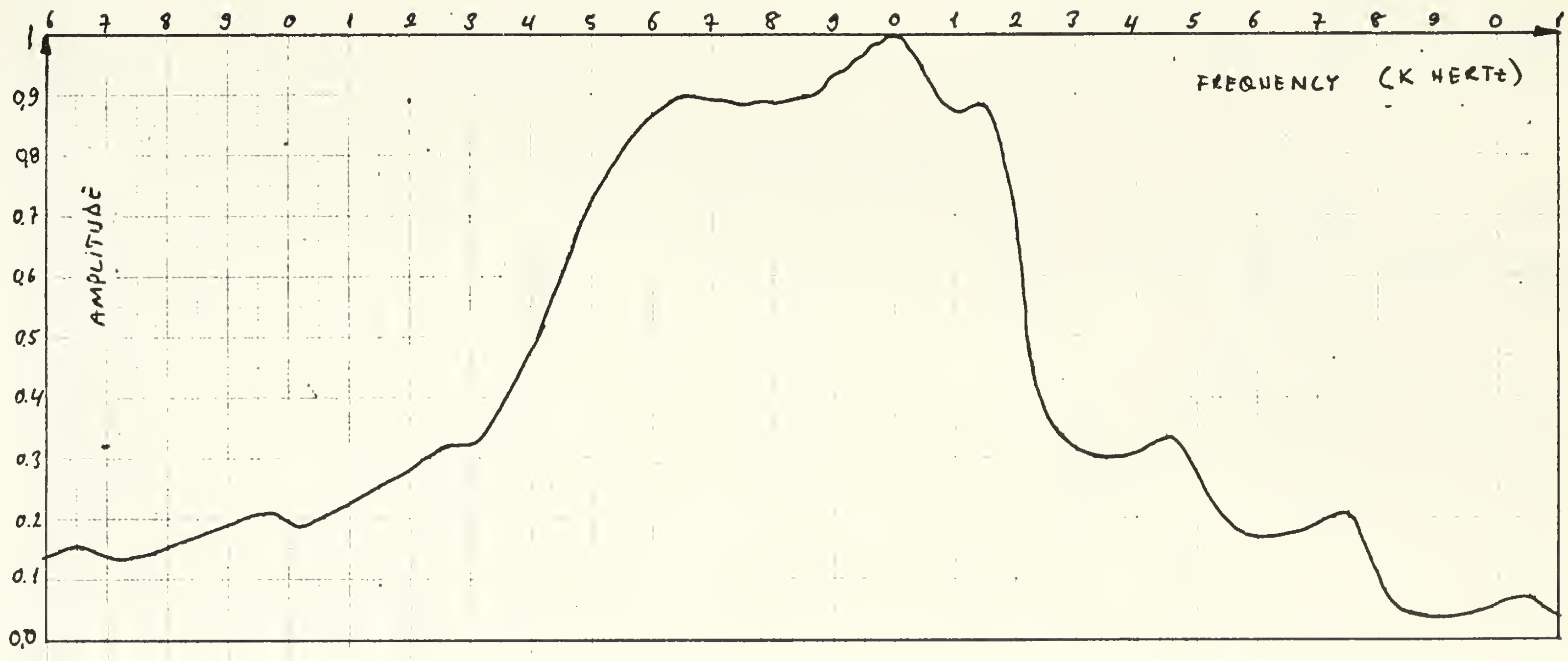


Figure 19 (b) The effect of the width of the input pulse signal on the IF filter
Figure 19(b) shows the same information on a linear scale for comparison.

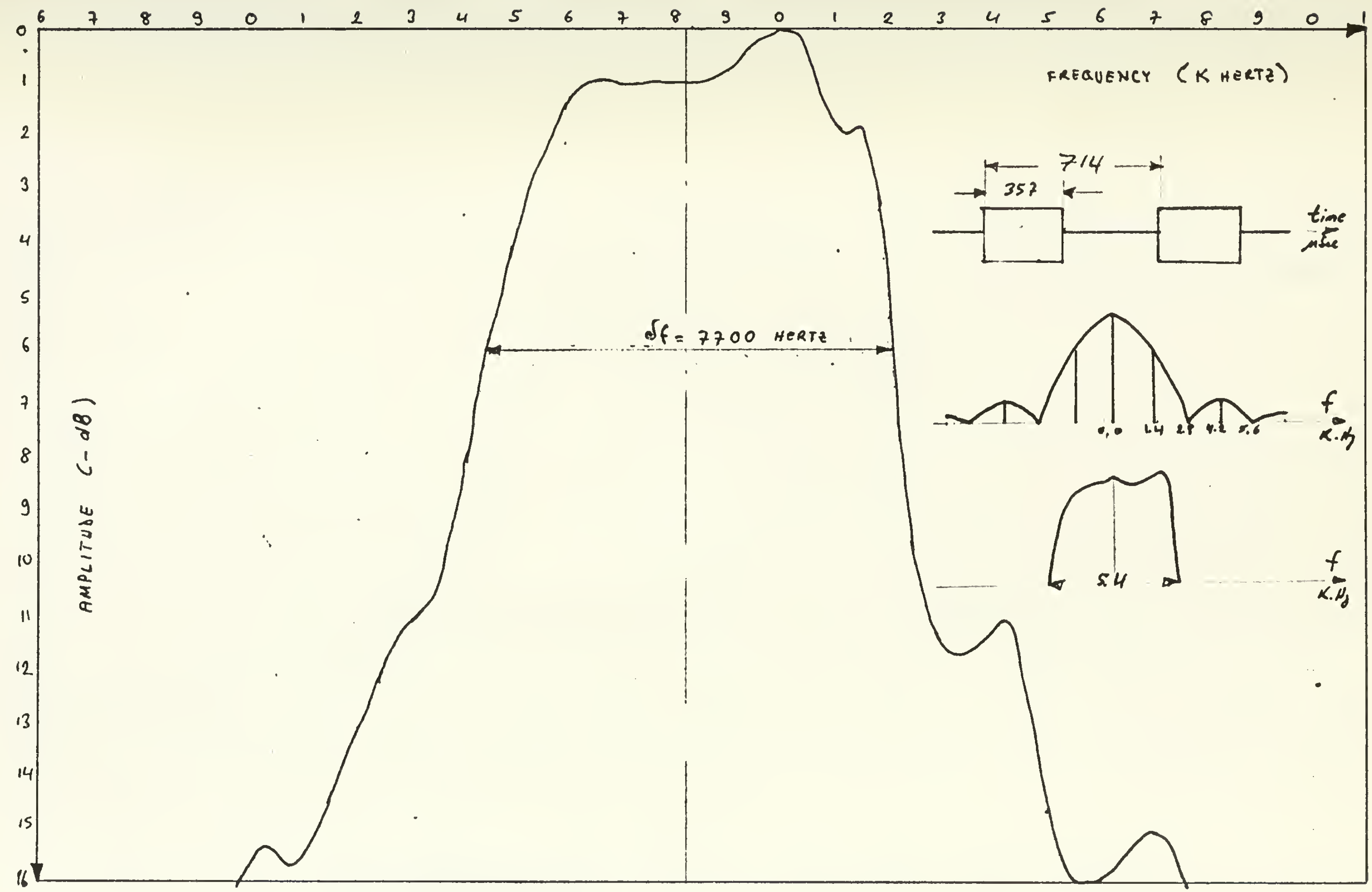


Figure 20 (a) The effect of the width of the input pulse signal on the IF filter
First sidelobe 11.1 dB lower than the mainlobe
Input pulse width τ = 357 μsec
Duty cycle = 1/2

Figure 20 (b) The effect of the width of the input pulse signal on the IF filter
Figure 20(b) shows the same information on a linear scale for comparison.

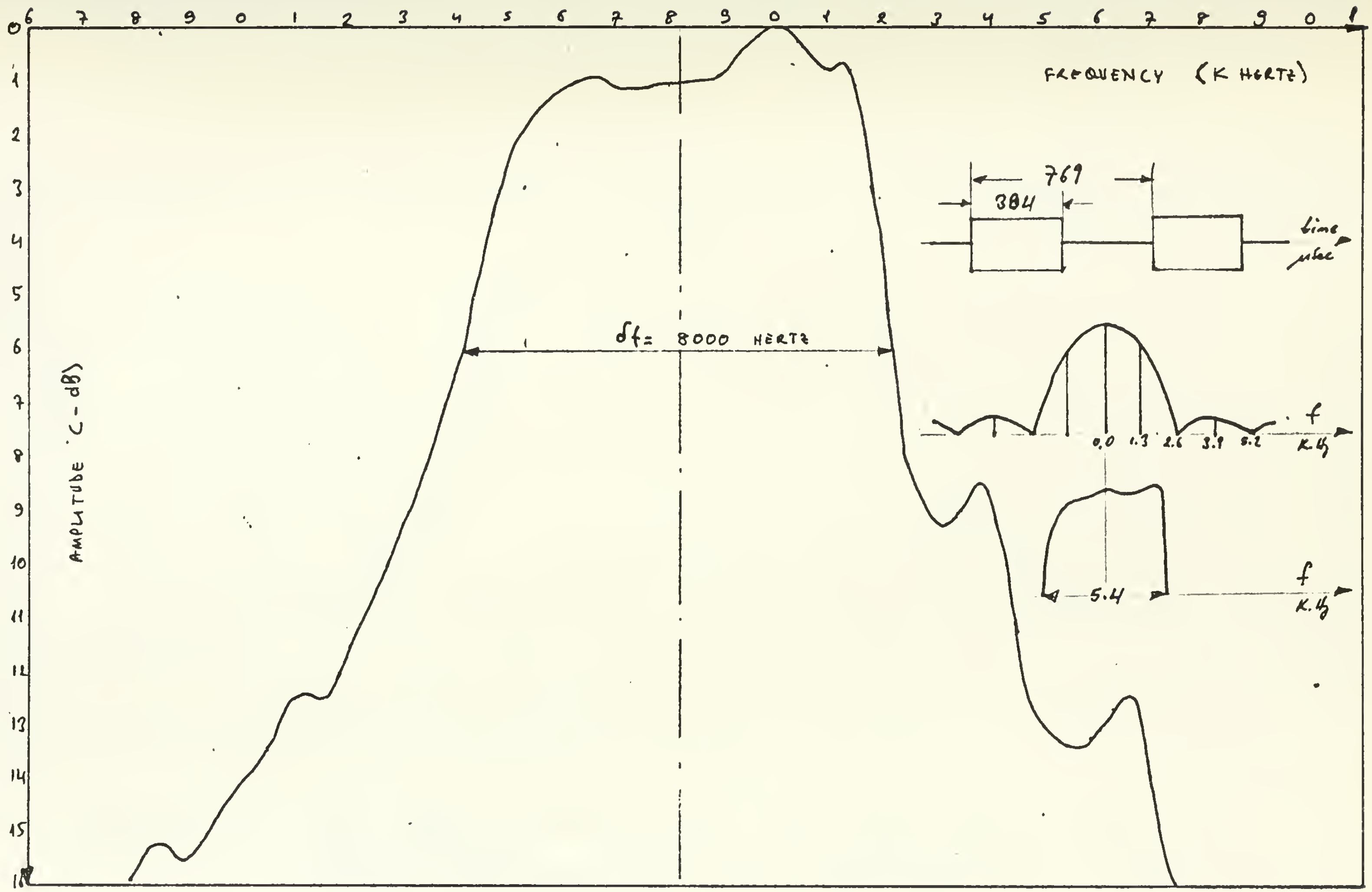


Figure 21 (a) The effect of the width of the input pulse signal on the IF filter
First sidelobe 8.4 dB lower than the mainlobe
Input pulse width τ = 384 µsec
Duty cycle = 1/2

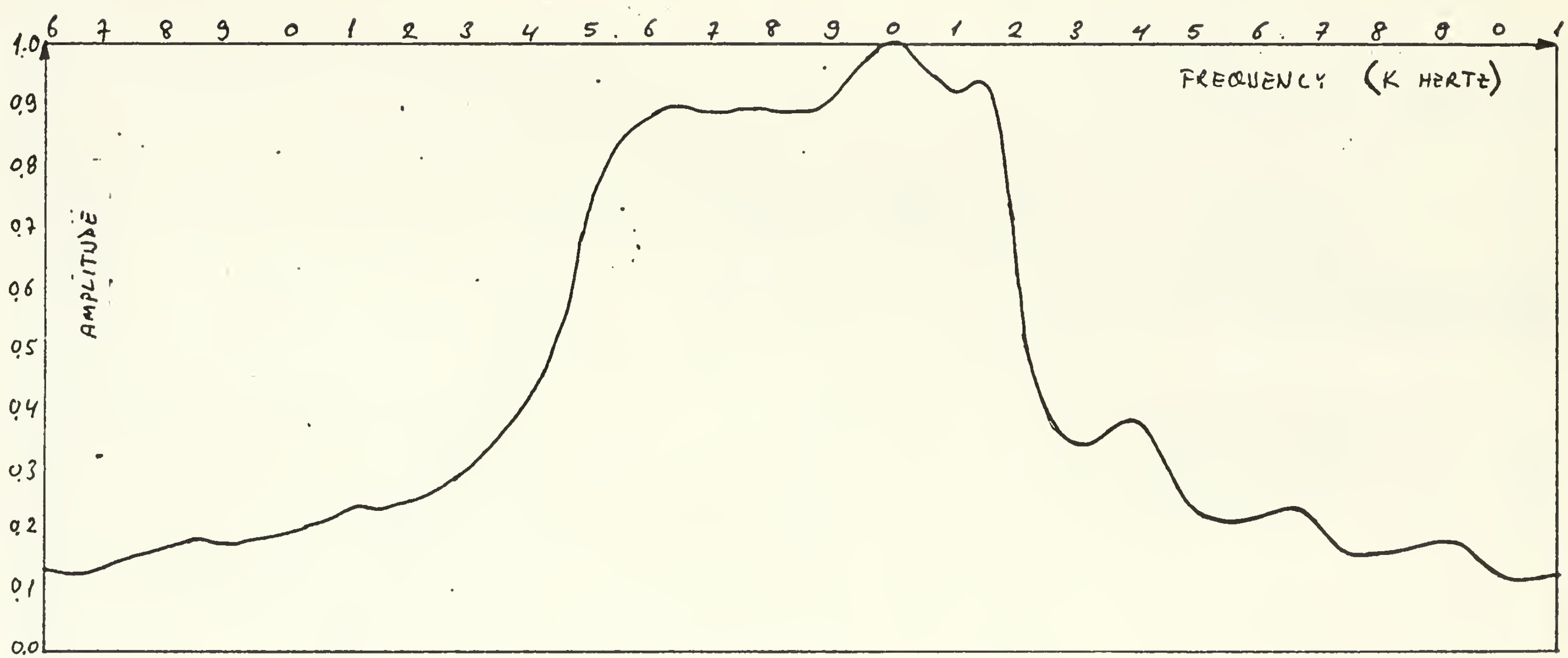


Figure 21 (b) The effect of the width of the input pulse signal on the IF filter
Figure 21(b) shows the same information on a linear scale for comparison.

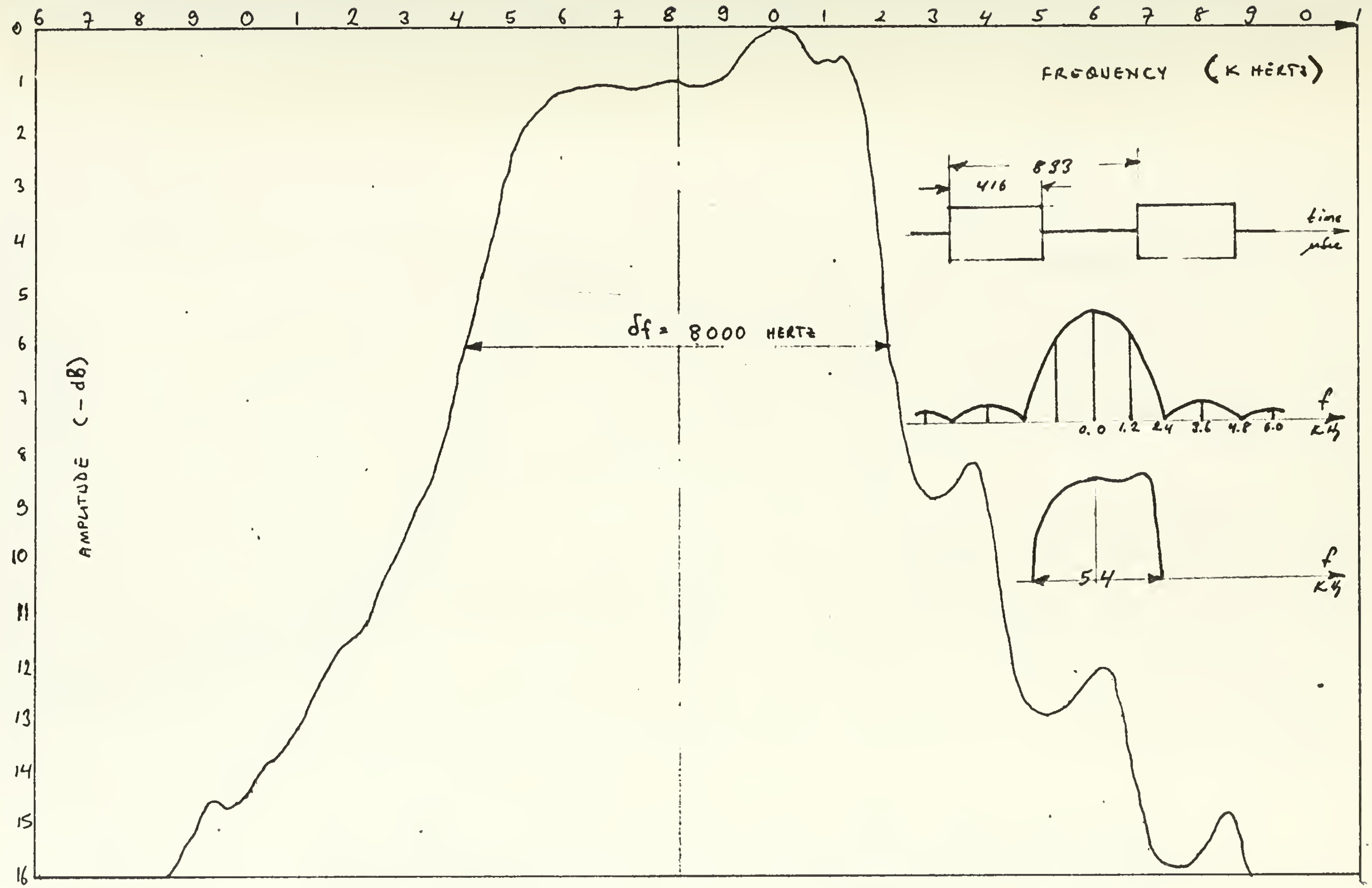


Figure 22 (a) The effect of the width of the input pulse signal on the IF filter
First sidelobe 8.2 dB lower than the mainlobe
Input pulse width τ = 416 μsec
Duty cycle = 1/2

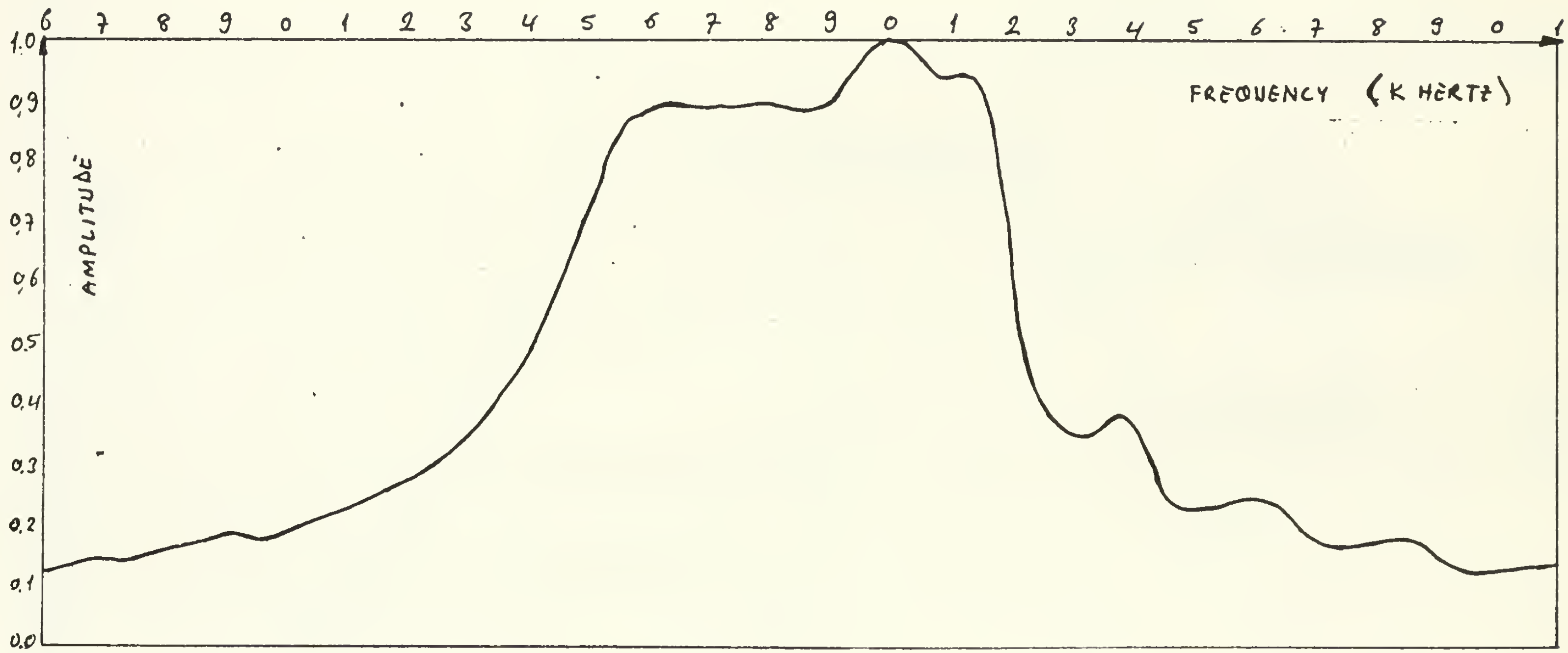


Figure 22 (b) The effect of the width of the input pulse signal on the IF filter
Figure 22(b) shows the same information on a linear scale for comparison.

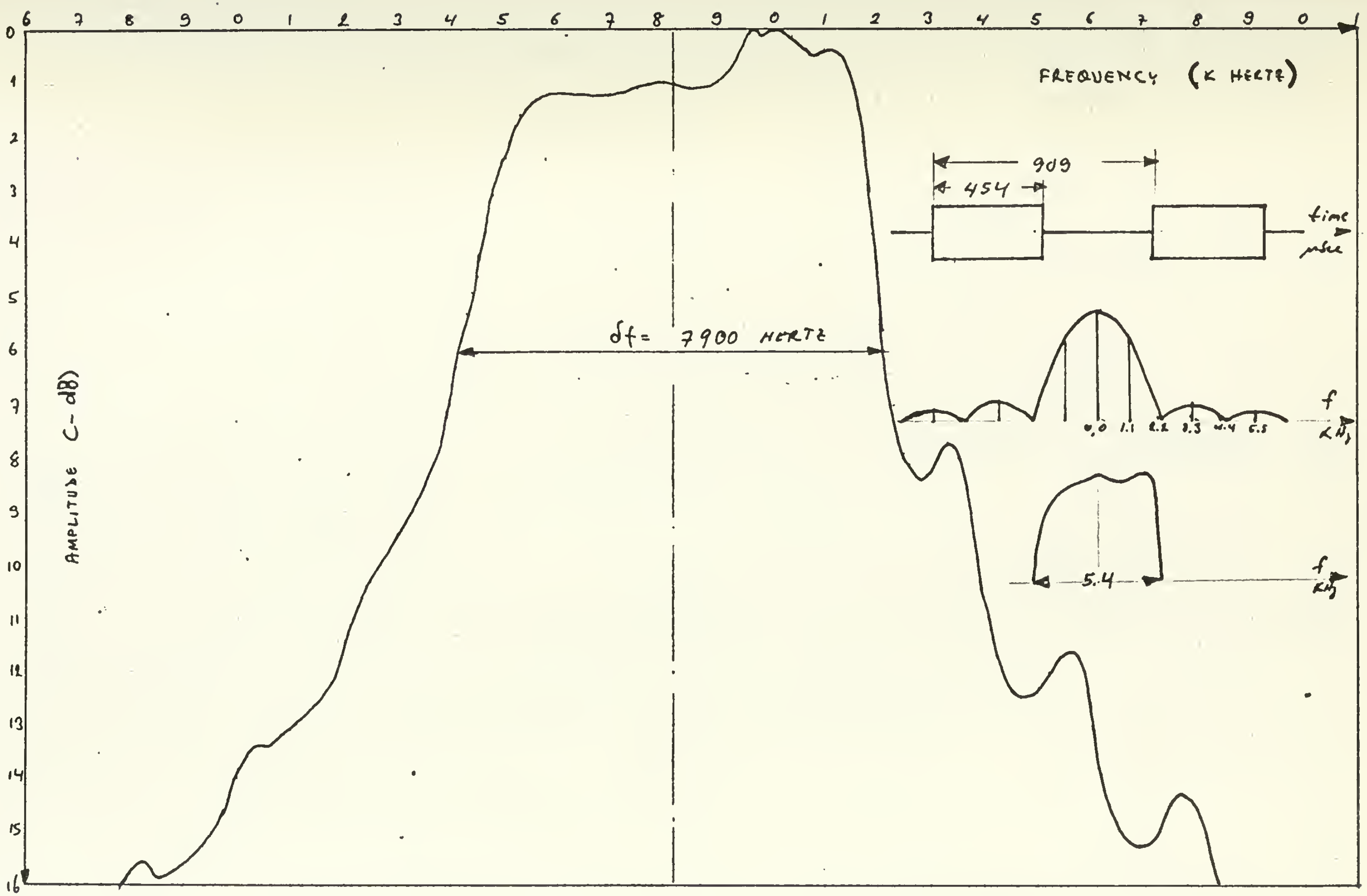


Figure 23 The effect of the width of the input pulse signal on the IF filter
First sidelobe 7.7 dB lower than the mainlobe
Input pulse width τ = 454 μsec
Duty cycle = 1/2

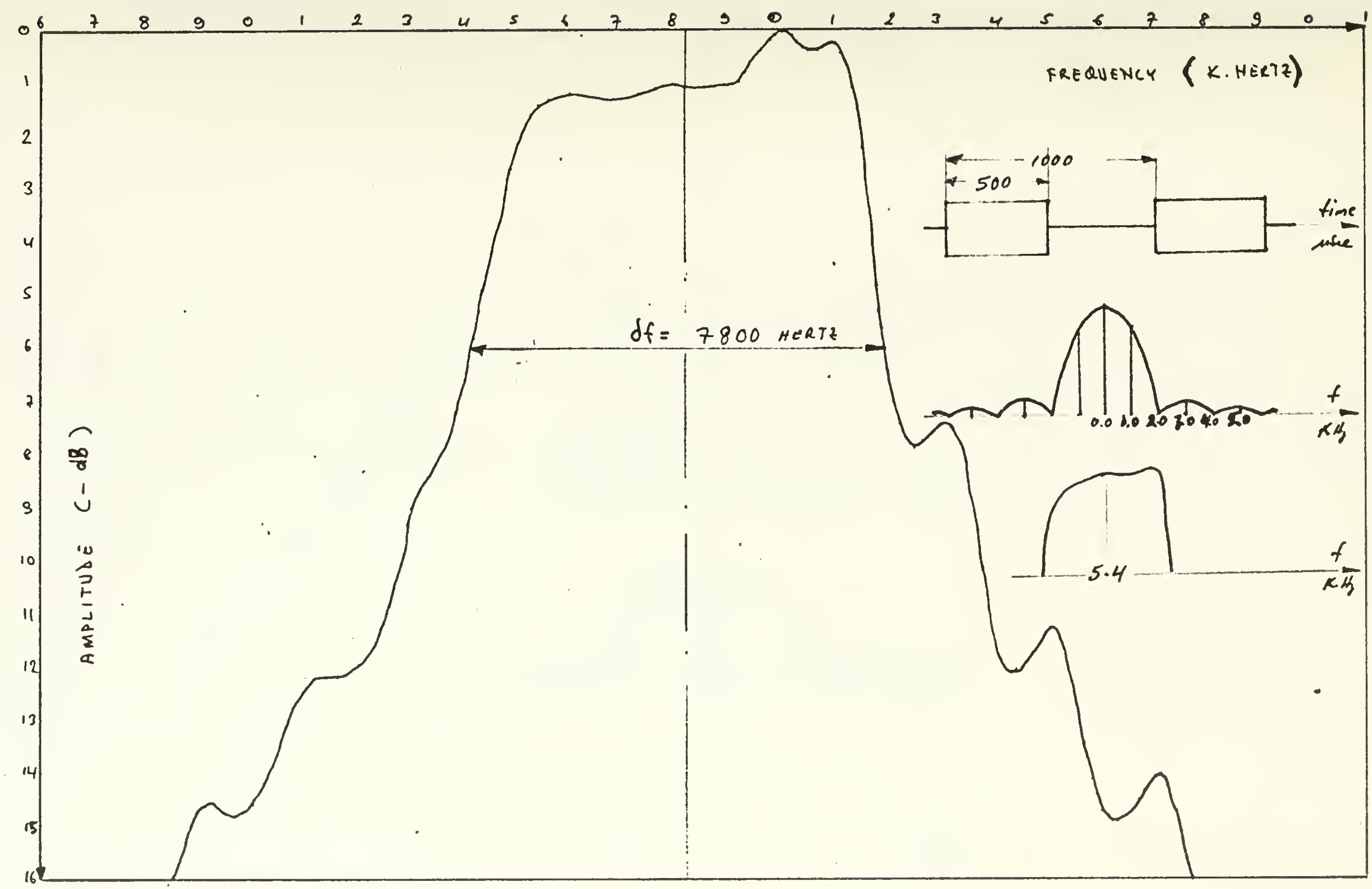


Figure 24 The effect of the width of the input pulse signal on the IF filter
First sidelobe 7.4 dB lower than the mainlobe
Input pulse width τ = 500 μsec
Duty cycle = 1/2

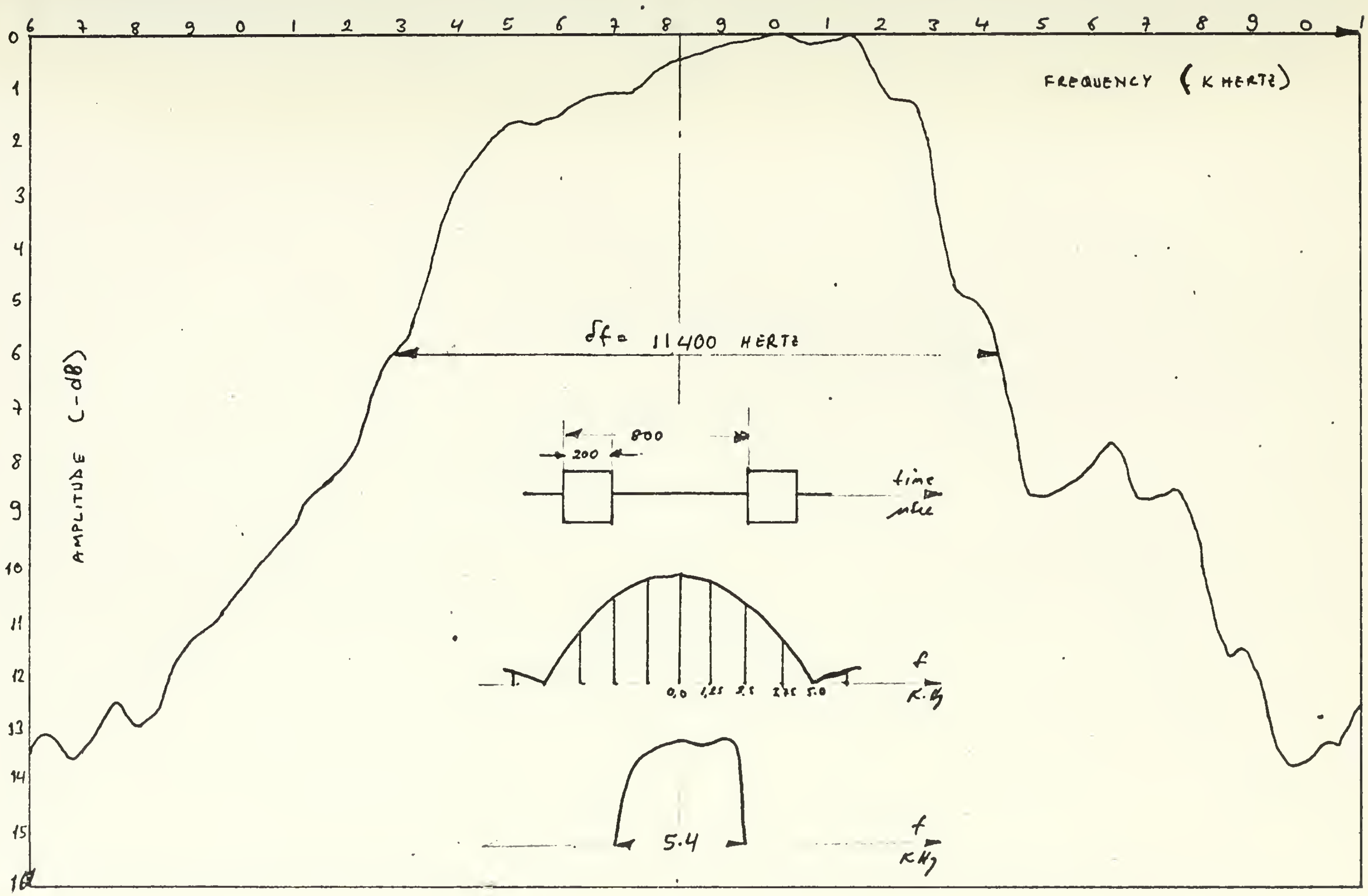


Figure 25 The effect of the width of the input pulse signal on the IF filter

First sidelobe 7.7 dB lower than the mainlobe

Input pulse width τ = 200 μsec

Duty cycle = 1/4

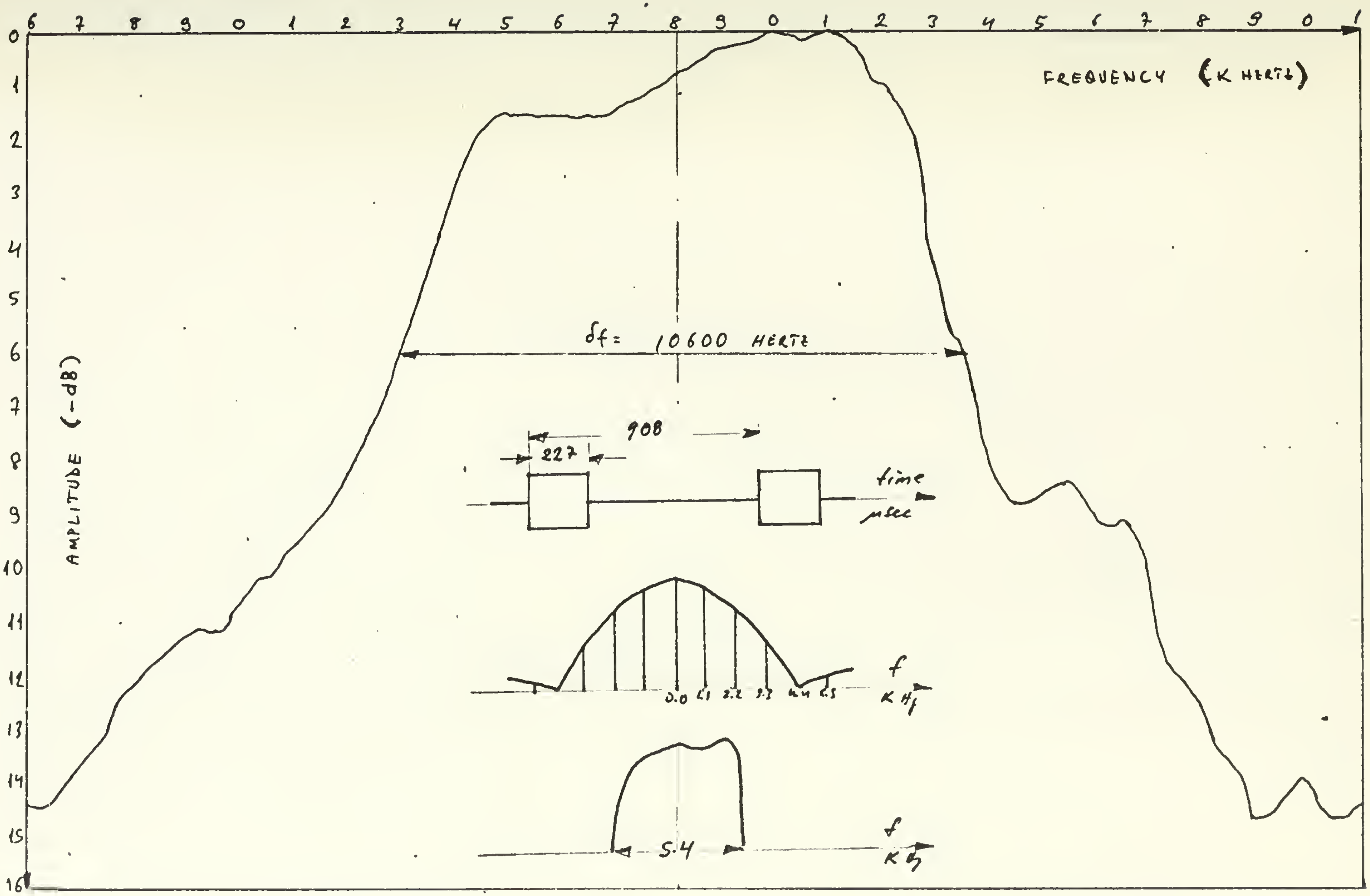


Figure 26 The effect of the width of the input pulse signal on the IF filter
First sidelobe 8.4 dB lower than the mainlobe
Input pulse width τ = 227 μsec
Duty cycle = 1/4

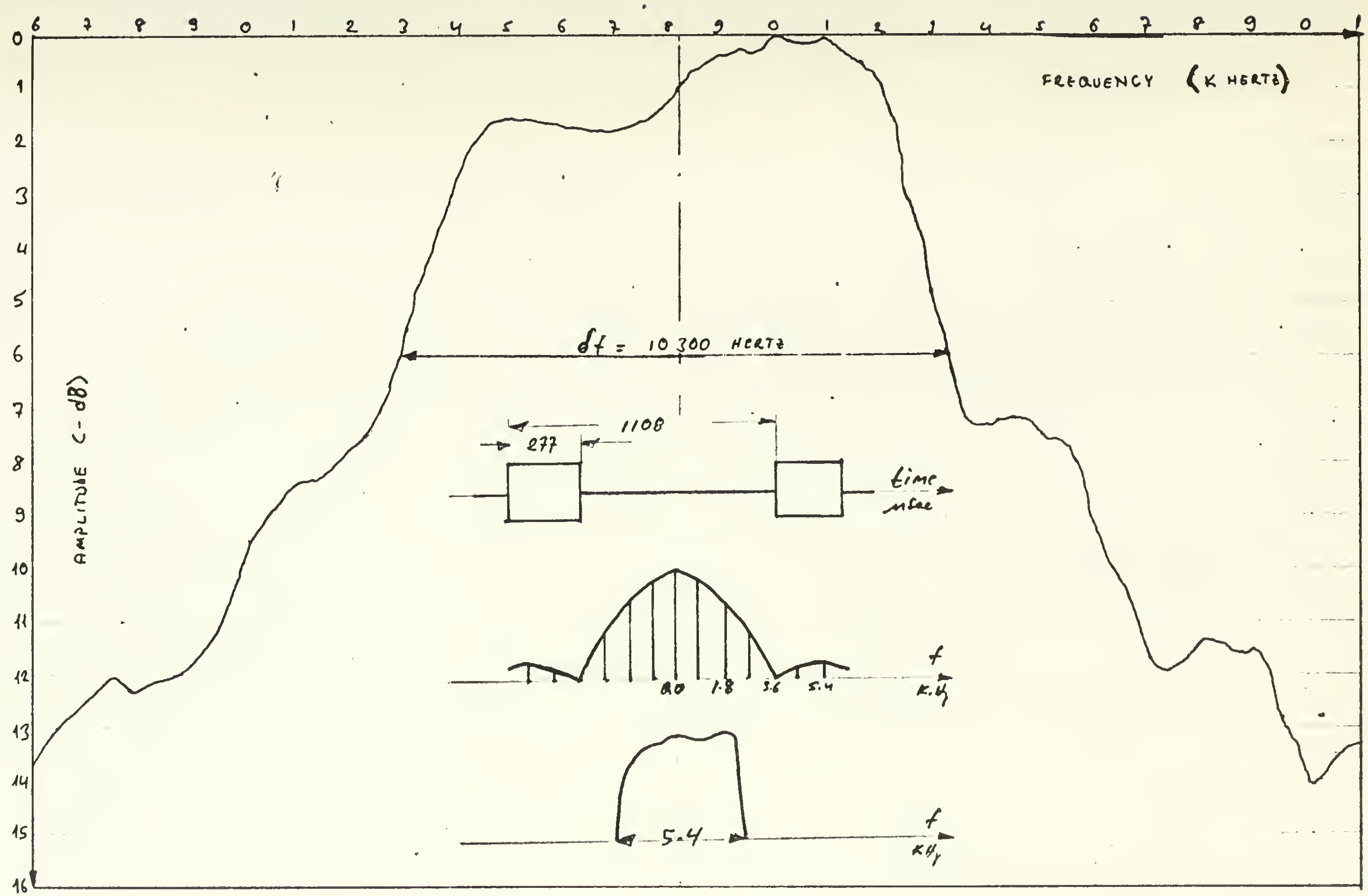


Figure 27 The effect of the width of the input pulse signal on the IF filter
First sidelobe 7.2 dB lower than the mainlobe
Input pulse width τ = 277 μsec
Duty cycle = 1/4

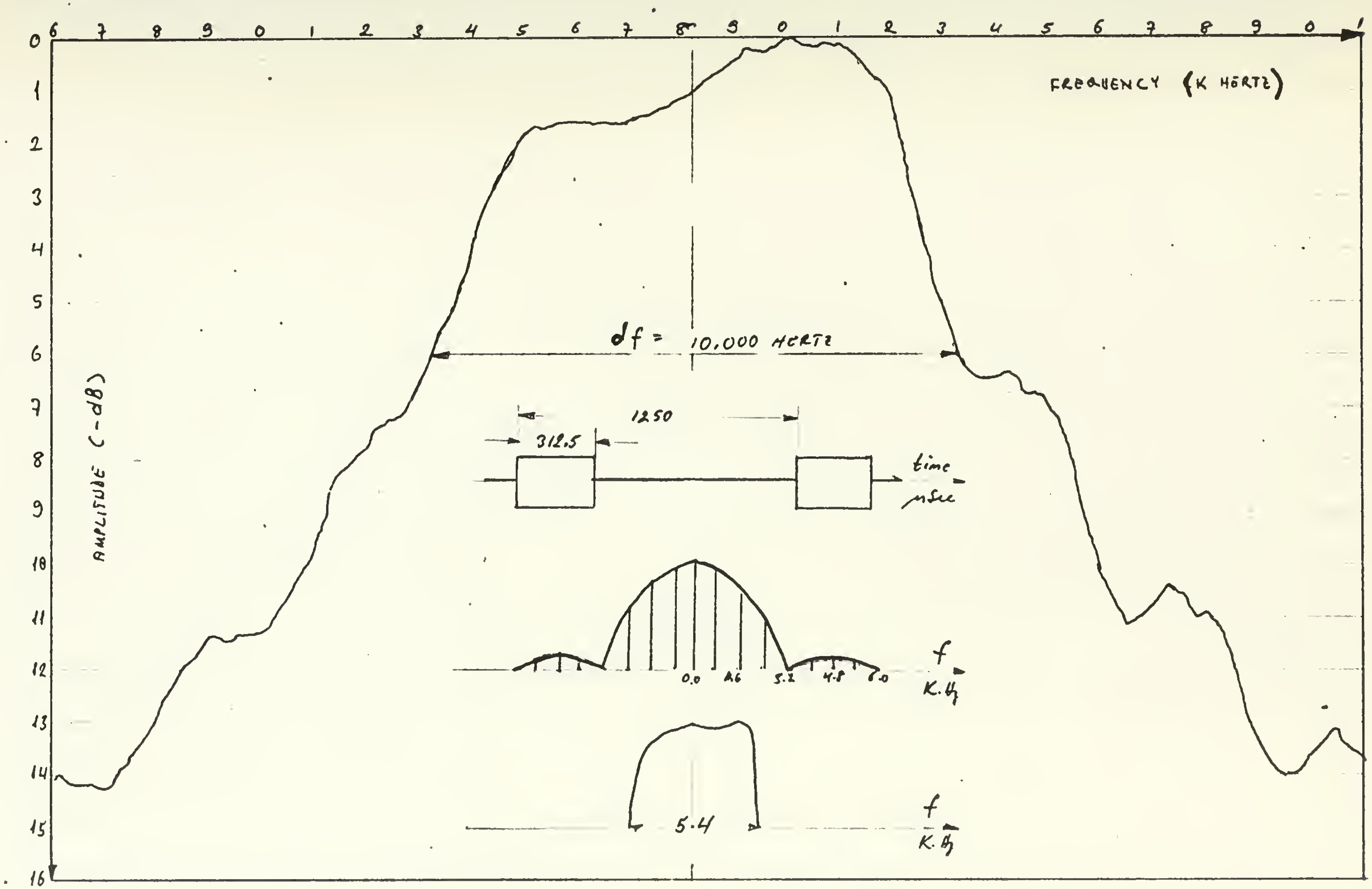


Figure 28 The effect of the width of the input pulse signal on the IF filter
First sidelobe 7.3 dB lower than the mainlobe
Input pulse width τ = 312.5 μsec
Duty cycle = 1/4

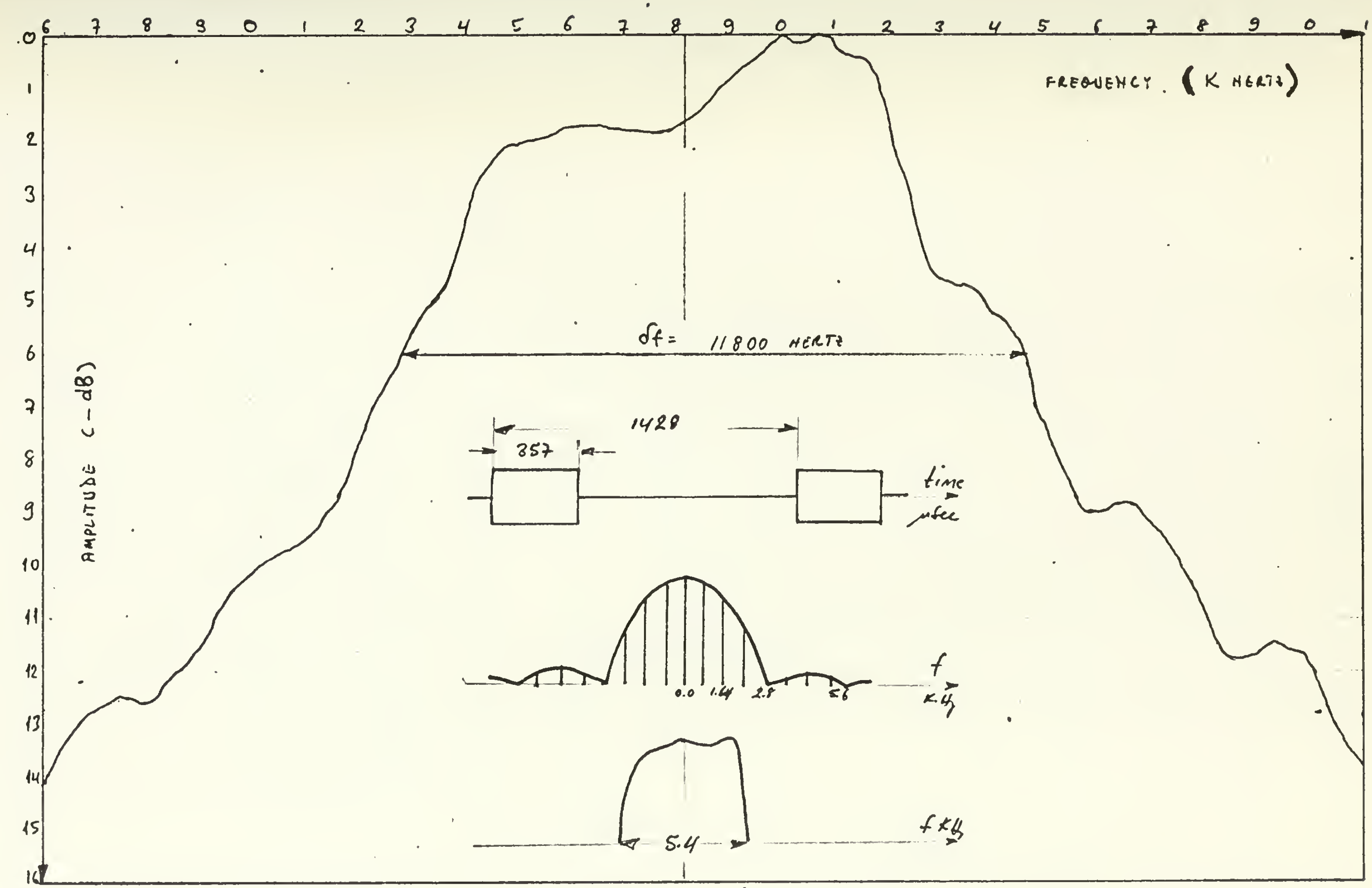


Figure 29 The effect of the width of the input pulse signal on the IF filter
First sidelobe 4.7 dB lower than the mainlobe
Input pulse width τ = 357 μsec
Duty cycle = 1/4

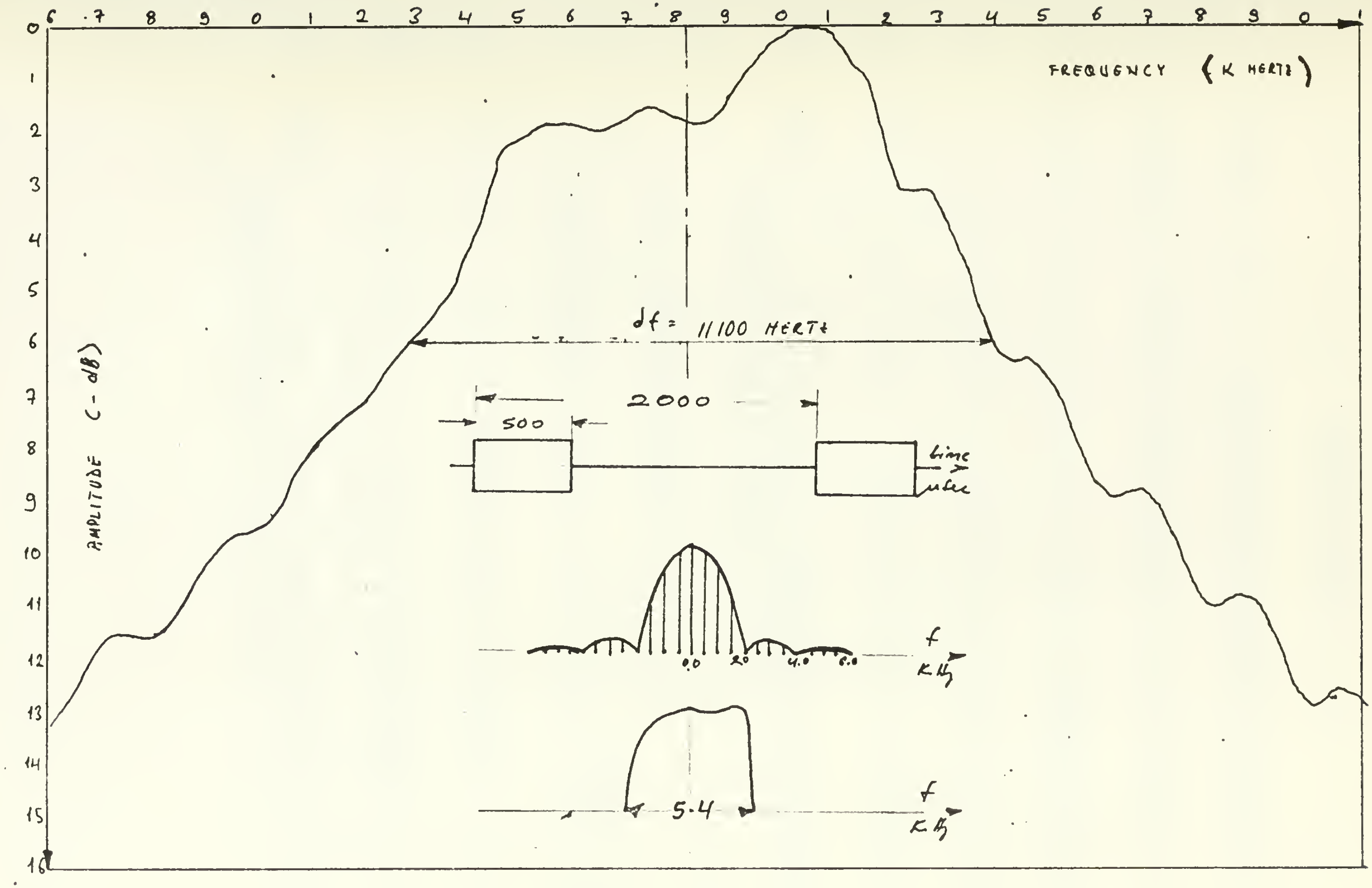


Figure 30 The effect of the width of the input pulse signal on the IF filter
First sidelobe 3.1 dB lower than the mainlobe
Input pulse width τ = 500 μsec
Duty cycle = 1/4

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